A SPREAD SPECTRUM COMMUNICATION SYSTEM

Cain Garrett

NAVAL POSTGRADUATE SCHOOL

Monterey, California

THESIS

A SPREAD SPECTRUM COMMUNICATION SYSTEM

by

Cain Garrett, Jr.

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This thesis reports on an investigation of a practical solution to the problem of providing covert, secure protection to the video channel of a remotely piloted vehicle (RPV) downlink. Spread spectrum technology employing acoustic surface wave devices and Golay complementary code sequences are utilized. Bandwidth reduction of the video data to a data rate of 200 kilobits per second is assumed.

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Block #19 Continued

ASWD
Remotely Piloted Vehicle
Video Data Link
Imagery Transmission
Transmitter/Receiver
Wideband Amplifiers
Processing Gain
Covert Communications

Block #20 Continued

A discussion of the need for such a system and the qualities that a successful system should possess is included in the introduction. Problems inherent in providing an adequate video link for military RPV usage are examined.

The thesis effort results in a transmitter/receiver combination with many of the desirable qualities of a video link for military RPV employment in a hostile environment. The hardware contains modifications that were made to facilitate system testing. Tests results are reported.

The emphasis of the thesis is on the satisfaction of the system specifications through hardware implementation.

A Spread Spectrum Communication System

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December 1974

ABSTRACT

This thesis reports on an investigation of a practical solution to the problem of providing covert, secure protection to the video channel of a remotely piloted vehicle (RPV) down-link. Spread spectrum technology employing acoustic surface wave devices and Golay complementary code sequences are utilized. Bandwidth reduction of the video data to a data rate of 200 kilobits per second is assumed.

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GLOSSARY OF SYMBOLS AND TERMS

Anti-jam	-	Capability of rejection of jamming signals.
ASWD	-	Acoustic Surface Wave Device.
BNC	-	Type connector for coaxial cable.
Communication Security	-	Message content coding.
CW	-	Continuous Wave.
Data Rate	-	Information transmission rate in bits per second.
DB	-	Decibel.
FH	-	Frequency Hop.
Golay Complementary Code	-	A pair of code sequences with an infinite correlation peak-to-peak ambiguity ratio when detected with a matched filter.
HF	-	High Frequency (3MHz - 30MHz).
Hybrid SS System	-	SS system employing two or more different methods of spectrum spreading.
Matched Filter	-	Filter whose impulse response is matched to the signal input.
MODEM	-	Modulator/demodulator pair.
PG	-	Processing Gain.
PN	-	Pseudo Noise.
RPV	-	Remotely Piloted Vehicle.
S-Band	-	2 Gigahertz to 4 Gigahertz Frequency range.
SMA	-	Type connector for coaxial cable.
SNR	-	Signal-to-Noise Ratio.
Spread Spectrum Techniques	-	Processes which cause the bandwidth of a signal waveform to be deliberately extended in order to achieve certain transmission advantages.

SS	-	Spread Spectrum.
Transmission Security	-	Protection against the disclosure of signal transmission.

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Thanks are also extended to Mr. James A. Kivett and Mr. Edward Koist of Hughes Aircraft Company, Ground System Division, Fullerton, California, for their assistance and cooperation during my industrial experience tour at their facility. Participation in the developmental and testing stages of their WCCM and PLRS spread spectrum systems gave much insight into and understanding of realizable spread spectrum systems.

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I. INTRODUCTION

The companion objectives of reliability, security, and speed have led to many changes and trade-offs in the concepts and design of military tactical and strategic communications systems. Although modern communications technology has reached a level of engineering sophistication that affords vast improvements in each of the major objective areas over our earliest communications networks, these improvements have been challenged by comparable improvements in signal jamming and signal interception techniques.

In general, the trade-offs have been attempts to balance the objective requirements to meet the particular employment of the communication network. A trade-off in a military communications system that enhances the attainment of one of the major objectives while degrading the system's ability to accomplish another major objective will always suffer since a hostile party will simply shift its emphasis toward the new weakness. A system that shows a major improvement in speed may have bandwidth and channel capacity requirements that make secure encryption difficult or hostile jamming a predictable threat.

Not all of the emphasis of the accomplishment of major objectives has been focused on the larger military communications systems. There exists many requirements for communications systems and networks that are small, even man-pack size, that must attain the same objectives of reliability, security

and speed. Here, as in the larger systems, hostile signal jamming or hostile signal interception of communication signals can be extremely detrimental to a military operation. Techniques that have become popular in larger communications systems such as high power (enhancing reliability) and elaborate encryption/decryption schemes (enhancing communication security) suffer in meeting the cost, size, and power consumption requirements of these smaller systems.

The frequency assignment problem becomes one of prime concern when the information to be transmitted is video data. Here the bandwidth before conventional encryption is sufficiently wide that this form of providing security becomes particularly impractical for general military usage and all but impossible for compact video systems.

A communication technique that promises simultaneous improvements in the reliability, security, and speed of military communications systems is spread spectrum communications. This technique involves the processing of a limited bandwidth signal of a specified average spectral power density to substantially increase the bandwidth of the information signal while decreasing the signal's average spectral power density over the transmitted bandwidth. This process results in a transmitted signal that has a negative signal-to-noise ratio over the increased bandwidth.

Whereas this reduced level of transmitted signal gives a measure of covertness to the transmission, the real security of the system comes from the coding involved in the signal spreading process.

System reliability is enhanced because of the inherent "anti-jam," "multipath rejection," and "interference rejection" that comes from the bandwidth spreading/compression scheme used in the communication MODEM.

There are currently three principle methods of spreading the spectrum of communications signals: pseudo-noise random code systems, frequency hop systems, and matched filter techniques, such as acoustic surface wave device (ASWD) matched filter systems.

In Psuedo-Noise (PN) systems, the information signal is combined with a noise-like pseudo random code. The resultant signal is one with a bandwidth that is approximately equal to that of the pseudo random code which is usually a thousand or more times as wide as the original information signal. The receivers of these PN systems must reapply a code identical to the one at the transmitter to compress the bandwidth of the transmitted signal and recover the original signal information. A major consideration in the design and implementation of PN spread spectrum systems is the maintaining of PN code synchronization between the transmitter and receiver.

Frequency Hop systems cause the information signal to be transmitted at discrete frequencies over a large frequency band at a random sequence and a rapid rate. The resultant waveform is one with a bandwidth that is equal to the range of frequencies through which the information signal "hops." The sequence of frequency transmissions are determined by a

pseudo random code in the transmitter. An identical code in the receiver causes the receiver to track the transmitter frequencies. As in the PN spread spectrum systems, frequency hop systems are plagued by the complexity at code synchronization problems.

Spread spectrum systems employing Acoustic Surface Wave Devices can avoid the necessity of maintaining PN code synchronization between transmitter and receiver by utilizing ASWD's in the system MODEM to provide a passive correlator for signal encoding/decoding and bandwidth expansion/compression. (Refer to Appendix A)

This thesis reports on the design, building, and testing of a spread spectrum communications system employing complementary coded acoustic surface wave devices.

A. STATEMENT OF THE PROBLEM

Current advances in the development of remotely piloted vehicles (RPV) have caused these airborne platforms to be considered by military planners for many uses formerly relegated to manned aircraft. Long on-station endurance time, low vulnerability and risk, optimum space utilization, and multiple launch sites are all factors which lead to the prediction that suitably equipped RPV's will play significant roles in future military surveillance operations as well as strategic plans for tactical offensive weapons exchanges.

Successful employment of these remote platforms require communications links that provide real-time information in a reliable, covert manner. The airborne equipment providing

the communications link must be small and lightweight to provide maximum space utilization for sensors and offensive weapons.

A particular problem has been that of providing a suitable down-link for the video data. Present experimental systems have either provided neither encoding nor processing gain for the video information or limited encoding without processing gain for the real-time video link. A hostile force would quickly exploit such system inadequacies, possibly leading to the failure of a critical mission.

The necessity of providing a down-link for real-time video data that is secure, covert, and reliable, and in a package that is suitable for tactical RPV employment, was the driving force for this thesis.

B. DEFINING THE THESIS

The initial consideration was that of determining how best to attack the problem of providing significant processing gain for a signal that already has a bandwidth that is several megahertz wide. As processing gain (PG), which is a measure of the anti-jam of the system, is given by

$$PG = \frac{\text{transmitted bandwidth}}{\text{information bandwidth}} ,$$

providing 30db of processing gain to a video signal of 2MHz bandwidth would require a spread spectrum bandwidth of 2GHz. State-of-the-art equipments with bandwidths in the gigahertz range are characterized by bulkiness and high power requirements.

The first breakthrough in a practical solution to the problem was a realization by Dr. Robert W. Means of the Naval

Underseas Center, San Diego, that a real-time video presentation could be provided from video data that has been reduced to a data rate of approximately 200 kilobits per second by redundancy reduction. (See Reference 1)

Since this video reproduction is adequate for many military applications, the objective of this thesis was to design, build, and test a system which could provide a lightweight spread spectrum down-link for 200 kilobit data.

1. General Considerations

The problem definition required that any solution must consist of a system which provides anti-jam rejection for the video channel. This requirement forced consideration of spread spectrum techniques. Since spread spectrum techniques involve ideas that are relatively new and equipment and devices that border on the very fringes of the state-of-the-art, the first consideration was whether this thesis should be an entirely theoretical investigation, a survey of current attempts to provide a practical solution, or an investigation leading to the design, building, and testing of a candidate solution.

Work done by Cocci in using acoustic surface wave devices with Golay coding (See Reference 2) to build a spread spectrum MODEM for voice communications provided encouragement and motivation for the consideration of having this thesis result in a working product capable of achieving the desired qualities of an RPV video down-link.

Further considerations led to the realization that providing a completed system capable of operating in the

S-band would not be necessary. In fact, a completed system with an output frequency centered in the HF-band would lend itself better to laboratory testing of the system's anti-jam capabilities. It was therefore decided that the finished product should demonstrate all of the aspects of a working system except operation in the S-band.

2. Physical Considerations

The constraint that the completed system should include an RPV transmitter that is size and weight limited was an obvious one. This requirement of compactness forced that integrated circuits be considered for maximum utilization. In addition to integrated circuits contributing to space efficiency they promised to be easily adaptable to the voltage levels expected on board the aircraft. Since many new concepts required circuits for which integrated circuits would not be available, these additional circuits were required to be carefully designed and constructed for space and power efficiency.

The system was to be designed to operate for long periods of time, for compatibility with the long endurance time expected of future RPV's.

An additional physical consideration of primary concern was that of mutual compatibility. The completed product was required to be capable of surviving conditions of extreme temperature changes as well as being able to operate effectively in a region of dense electro-magnetic activity.

Simplicity was a physical constraint that was forced by the other physical requirements. The direct approach usually resulting in fewer components and hence a more reliable system.

Cost being a factor, off-the-shelf parts were sought for their economy as well as their proven dependability, but when off-the-shelf items could not be provided with the requisite size and minimum power requirements, other options were dictated. So the overriding physical consideration was that of compactness followed by simplicity, cost economy, and power efficiency.

3. Theoretical Considerations

The requirement that the down-link for video data be covert and have anti-jam rejection meant that any acceptable system would require processing to spread the spectrum of the video information signal bandwidth. Since an assumption of covertness alone is not sufficient to guarantee adequate security, some type of information encoding had to be provided for as well. Theoretically the signal encoding could be obtained by the same mechanism that spread the spectrum of the signal. This method of approach was to be pursued where practicality and consistency with physical constraints permitted.

Although a reduction of the video data rate to 200 kilobits per second represented a significant decrease in the video signal bandwidth, providing high processing gain by band spreading alone would still require a spread spectrum bandwidth of several hundred megahertz. Theoretically,

processing gain could be increased over that predicted by the spread spectrum signal bandwidth to information signal bandwidth ratio if additional processing gain can be provided from the coding/decoding method. (See Reference 3 and Appendix B.)

Since the ASWD matched filters reported on in Reference 2 and described in Appendix A were available and since they had demonstrated the possibility of providing a spread spectrum signal by using Golay complementary coded devices, it was decided that these units should be given every consideration as candidates for inclusion in the final product.

It was also decided that if these devices were used in the final product that comprehensive testing should be performed to determine if the method of coding did in fact contribute to the processing gain of the system.

Section II gives a detailed discussion of the system specifications. Sections III and IV address the design and construction portion of the thesis project with major emphasis on why one method of approach was chosen in preference to another. Section V is a description of the testing procedure of the completed product, and Section VI reports on the results of testing the system. Section VII summarizes the thesis effort. Appendices are provided that supplement and illuminate the main text.

II. DESIGN SPECIFICATIONS AND OBJECTIVES

The overall objective of the system design was to provide a secure, covert communications link for the transmission of reduced video data from a remotely piloted vehicle (RPV) to a surface based command center.

Since the system transmitter was to be part of the payload of a weight and power limited RPV, it was required that the transmitter be as light as possible (with a goal of one pound exclusive of power supply) and have a minimum power budget commensurate with range requirements.

The system was to be capable of surviving a maximum doppler rate of 50 Kts, and to be capable of communications at a maximum of 100 nautical miles.

At a reduced video data rate of 200 kilobits per second, the system was designed to provide a minimum of 20db of anti-jam with a goal of 25 to 30db of anti-jam.

The system was to employ spread spectrum techniques utilizing an acoustic surface wave device modem for signal bandwidth expansion/compression and data encoding/decoding.

Although the system is expected to be operated in the S-band, the model designed, built, and tested for this thesis study was designed to operate in the HF band to facilitate laboratory testing; conversion of the system to S-band operation being a trivial extension of the thesis design.

In addition to the transmitter of the system having a weight specification of approximately one pound, both the

transmitter and receiver were required to be compact in order to economize on space and to facilitate portability. And although no specific restrictions were imposed on the receiver size, the receiver was to be made as small as the state-of-the-art would permit commensurate with transmitter size specifications.

Although the system was not required to provide a power supply for the RPV transmitter, the transmitter was required to be operable using standard aircraft voltage levels.

All components were required to be shielded in such a manner to permit satisfactory operation in an environment of high electro-magnetic activity and to be constructed using engineering practices that would minimize spurious radiations.

Test points were to be provided where necessary to permit testing of the system's performance in various noise and interference environments.

III. SYSTEM DESIGN AND FABRICATION

Consistent with the size and weight specifications defined in the earlier sections of this thesis, every consideration was given to space and weight economy in the design phase of this effort. The first major space saving option occurred in the consideration of a system waveform. Whereas other system specifications had already imposed the requirement of a spread spectrum waveform, the many candidate methods of providing spread spectrum modulation, as listed in Section I, left the choice of a particular spread spectrum waveform a significant option of consideration.

Since the choice of a spread spectrum waveform dictates much of the rest of the system design, the features of the various methods of providing spread spectrum modulation which led to the final choice will be discussed in detail.

A. CHOICE OF SPREAD SPECTRUM MODULATION SCHEME

Four candidate methods of providing a spread spectrum waveform were considered. These included pseudo noise (PN) code modulation, frequency hop (FH) modulation, a hybrid combination of PN code modulation with frequency hopping, and modulation by acoustic surface wave device. Detailed discussions of these methods of spread spectrum modulation are included in References 2 and 5. Only the salient features of each method will be discussed here in relation to the impact of these features on the final design choice.

Refer to Figure 1 for a simplified block diagram of a PN code spread spectrum system.

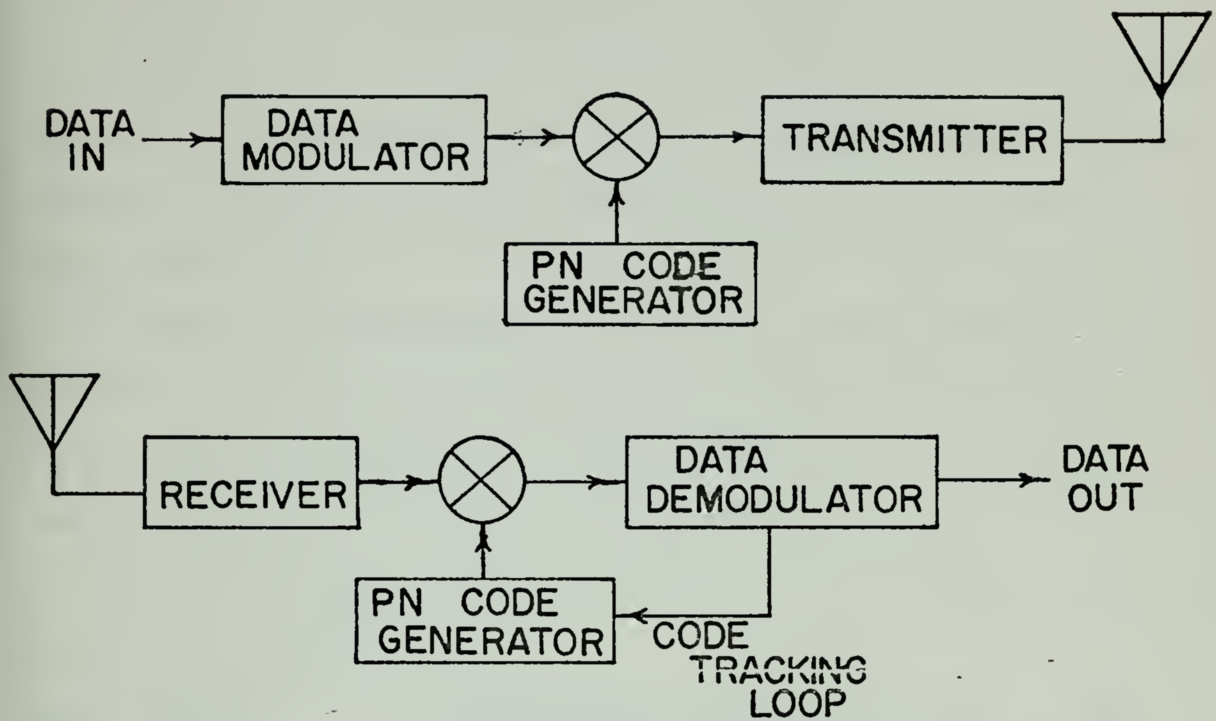

Figure 1. PN Code Spread Spectrum System.

The input data receives a bandwidth expansion when the data signal is combined with a PN code whose bandwidth is very large compared to that of the information signal. The resultant spread bandwidth is essentially that of the high rate PN code. The output signal is therefore characterized by a wide bandwidth and a low average spectral power density. At the receiver a PN code identical to that of the transmitter and locked to the transmitter code's phase is applied to the spread spectrum signal to compress the bandwidth of the data, thereby recovering a replica of the original data. Without dwelling on involved system specifics, (interested

readers should refer to Reference 5) the advantages and disadvantages of a PN code system with respect to its use in the final design will be listed. Features common to all candidate systems will be omitted.

Relative advantages of a PN code spread spectrum system include: (1) accurate ranging, (2) superior communications security capability, (3) multiple access, and (4) multi-path rejection.

Relative disadvantages of a PN code spread spectrum system include: (1) slow code acquisition in the receiver, (2) complicated PN code tracking system, and (3) sophisticated (space consuming) PN code generator required.

Refer to Figure 2 for a simplified block diagram of a FH spread spectrum system.

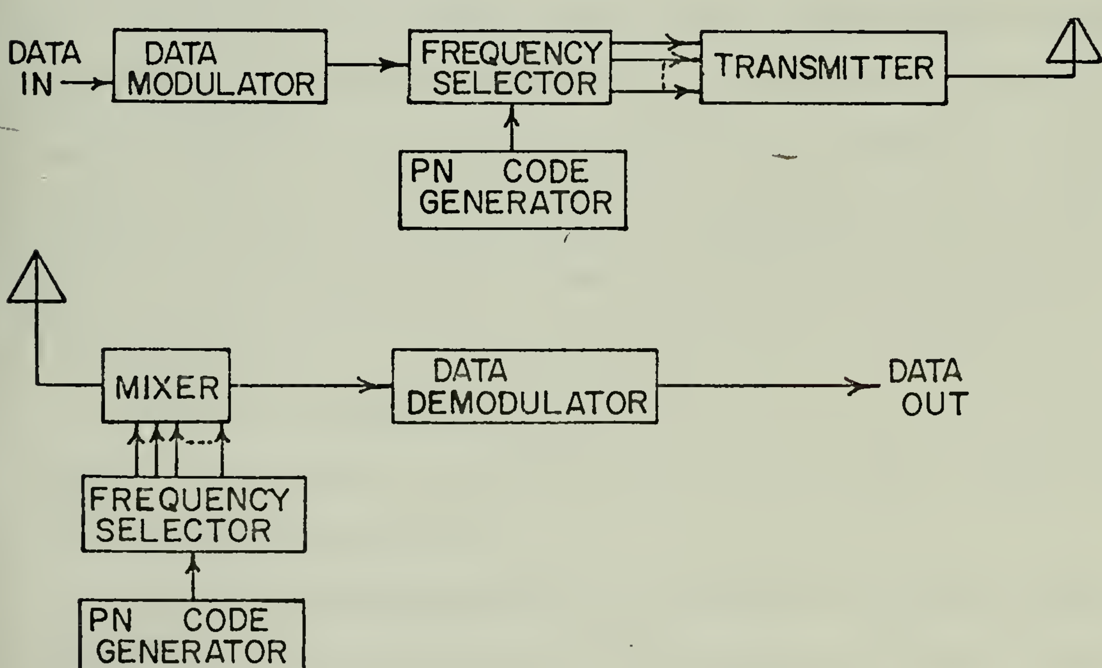

Figure 2. FH Spread Spectrum System.

In the FH method of providing a spread spectrum waveform, the data signal is transmitted at various frequencies throughout

a given frequency range in a pseudo random manner. The frequency of transmission is determined by a PN code generator which causes the transmitted signal to "hop" through the spread bandwidth, dwelling at each discrete frequency for a time that is determined by the chip length of the PN code. The resultant transmitted waveform is one with a bandwidth that is equivalent to the frequency range "hopped" through and with an average spectral power density that is significantly below the ambient noise levels. As in the PN code system, the receiver must have an exact replica of the transmitter code, in perfect code synchronization so that the receiver will always be "looking" at the proper frequency slot and will thereby compress the spread spectrum signal to a replica of the transmitted data. (Note that since processing gain is given by the ratio of the spread bandwidth to the input data bandwidth, the PN code generator need not operate as fast as that of the PN code system for a given level of processing gain. Hence, the PN receiver code tracking problems in the FH system are not as severe.)

Relative advantages of a FH spread spectrum system include: (1) faster code acquisition, (2) easier code tracking, (3) multiple access.

Relative disadvantages of a FH spread spectrum system include: (1) complicated PN code tracking system, (2) frequency synthesizer required, (3) relatively large system, and (4) only moderate ranging capability.

Hybrid systems containing both PN code spread spectrum modulation can take many forms. A representative block

diagram of such a combination is given in Figure 3. The particular system design is predicted by the features of

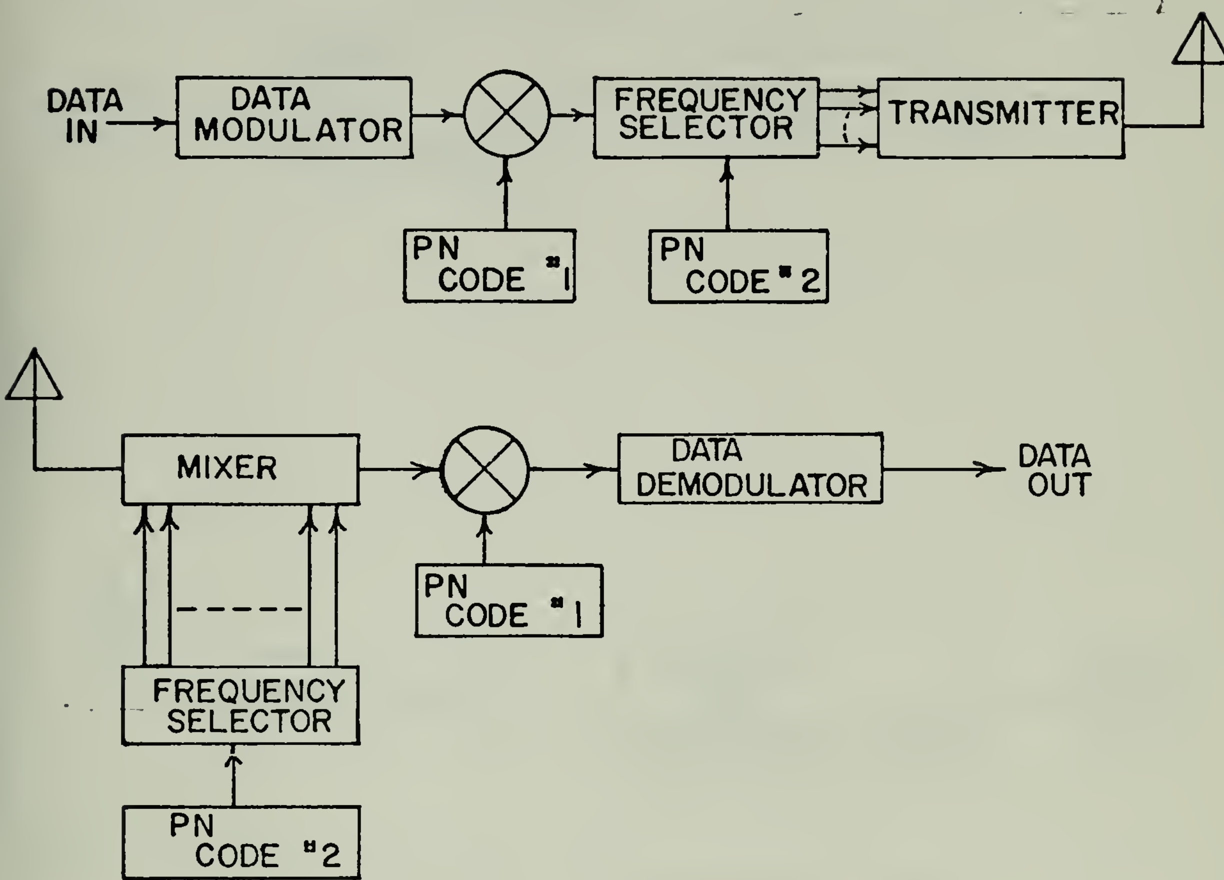

Figure 3. Hybrid FH/PN Code Spread Spectrum System.

the separate modulation methods that are to be obtained in the combination system. An advantage of significant interest is the larger processing gain that can be obtained using specified code rates. This improvement results from the fact that the output bandwidth is spread by two different methods and the effect of the combination is multiplicative. The disadvantage that prevails here is the sophistication of the PN code synchronization problem becomes doubled. This sophistication mainfests itself in larger system sizes and complexities.

See Figure 4 for a simplified block diagram of a representative ASWD spread spectrum system.

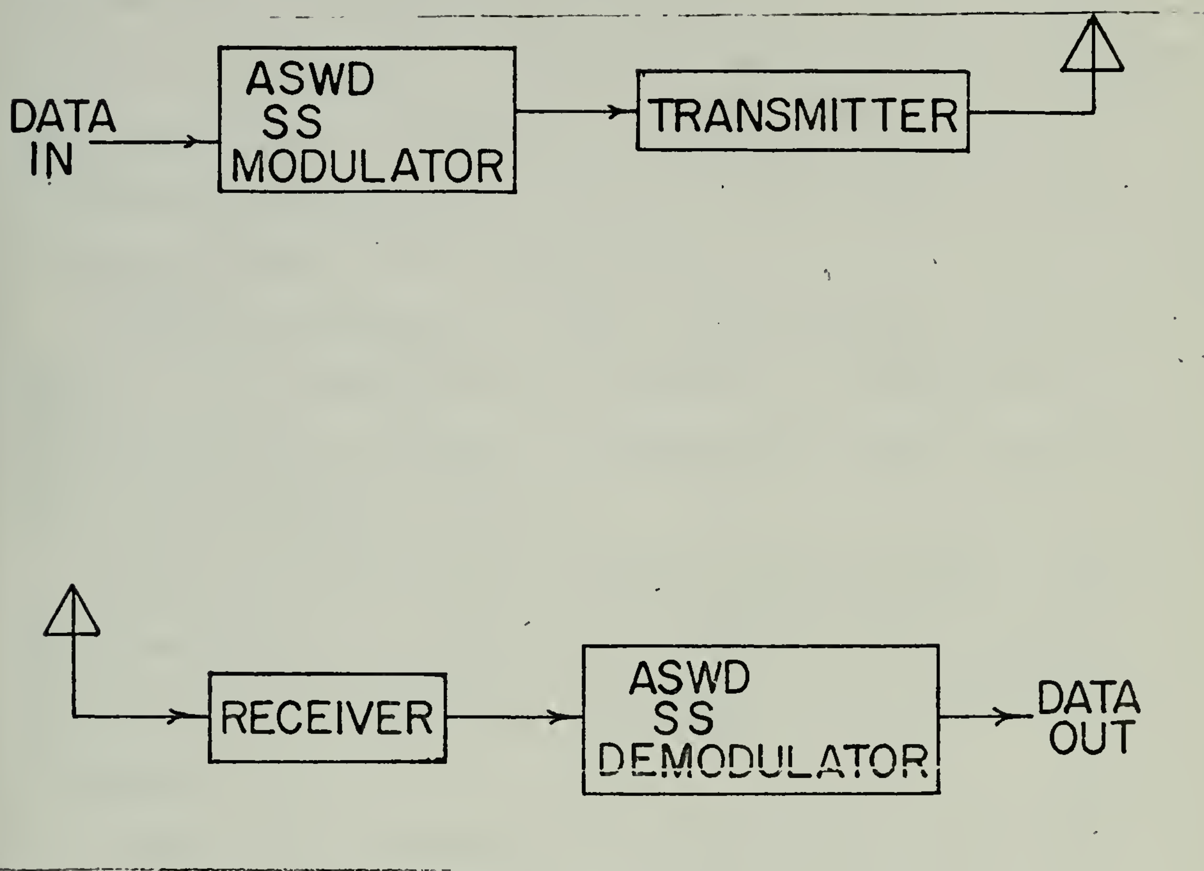

Figure 4. ASWD Spread Spectrum System.

The input data is caused to excite coded transducers of an ASWD in such a manner as to produce an output signal with a spectrum that is spread by an amount that is a function of the ASWD code length and the physical characteristics of the device. (Refer to Appendices A, B, and C, and to References 2 and 5.) The receiver contains an ASWD matched to that of the transmitter which forms a passive matched filter for the received spread spectrum system, collapsing the signal bandwidth and recovering the transmitted information. (An extension of the system represented by the basic block diagram

of Figure 4 includes systems containing multiple devices, such as the ones reported on in Reference 2 and in this thesis.) Of noteworthy interest is the fact that the system of Figure 4 does not contain a PN code generator. This significant advantage in simplicity is not without a price as will be seen in later discussion.

Relative advantages of ASWD systems include: (1) simplicity, (2) insensitivity to doppler, (3) reduced code synchronization problems, (4) comparative small size, and (5) passive matched filter.

Relative disadvantages of ASWD systems include: (1) low communication security, (2) limited processing gain available, (3) no ranging capability, and (4) unproven technique.

Along with advantages and disadvantages of the candidate systems that resulted from natural properties and characteristics of the components of the various systems, another important aspect of the individual systems had to be considered. The factor alluded to was the availability of critical system components. Since this thesis effort was to result in a product capable of being laboratory tested, due concern was warranted for the availability of devices and parts necessary to complete the chosen type of system early enough for testing to be completed within the time allowed for the thesis research.

After consideration of the advantages and disadvantages of the candidate spread spectrum methods, the apparent ideal choice was that of a hybrid system containing a PN code

spread spectrum waveform that "hops" through a block of frequencies at a rate determined by a different PN code pattern. This arrangement would permit maximum processing gain with communication security as well as transmission security. However, system complexity, projected system size, and the fact that critical system components were not available forced the rejection of this hybrid proposal. A similar argument forced a rejection of the apparent second choice which was to use a spread spectrum waveform generated by the PN code method. Mr. James A. Kivett and other engineers at the Hughes Aircraft Company, Ground Systems Division, Fullerton, California, were extremely helpful in many discussions of the availability of the companion components of the candidate systems.

The next choice, and the ultimate one, was that of using an ASWD waveform for the spread spectrum modulation. Although this choice was largely dictated by the fact that two Golay complementary coded pairs of ASWD matched filters were available, this choice did not represent an unacceptable concession. Aside from the advantages listed in earlier paragraphs, providing the spread spectrum waveform via Golay coded ASWD pairs promised interesting features for investigation. One such feature for investigation was the expectation that the Golay coding would produce a measure of processing gain greater than that predicted by a simple comparison of the information bandwidth with the spread spectrum bandwidth. These advantages and special features along with the opportunity of working with a device that

bordered on the very fringe of the state-of-the-art provided special stimulus for the design work of this thesis.

B. ELEMENTS OF DESIGN

The choice of the Golay coded ASWD's as sources of the spread spectrum waveform imposed a requirement that the transmitter/receiver combination include some form of multiplexing. Two separate channels of spread spectrum information result from the complementary coding. A second design problem was that of modulating the ASWD pair with a data rate of approximately 200 kilobits per second. The devices had already been successfully utilized in a voice MODEM, (see Reference 2), but certain modifications to the modulation scheme would be necessary to make the devices compatible with this higher rate of information transmission. Other design problems included amplification and transmission of the multiplexed signal, reception and amplification of the radiated signal, demultiplexing and demodulation of the received signal, and system integration for space economy and transmission efficiency.

So the design problem can be divided into the following elements: (1) modulating the ASWD's, (2) multiplexing the complementary waveforms, (3) amplification and transmission of the multiplexed signal, (4) reception and amplification of the spread spectrum signal, (5) demultiplexing and demodulating the received signal, and (6) system integration.

Each element will be discussed only in regard to its impact on the final product. Specific circuitry will be discussed only when the approach is significantly different

from normal practice, or when a cause and effect relationship is not obvious. The system integration will be covered in a separate section.

1. Modulating the Acoustic Surface Wave Devices

As described in Appendices A and B, an ASWD yields spread spectrum waveforms centered about the resonant frequency of the device when it is excited by a pulse. Cocci noted in his work, reported on by Reference 2, that optimum response occurs when a pulse of RF with the appropraite width and amplitude is used to excite the device. The pulsed RF should have a frequency that is approximately equal to the resonant frequency of the device. Given that the frequency of the pulsed RF should be 21.4 megahertz, it remained to determine the optimum width and amplitude characteristics of the modulating pulse. Examination revealed that the ASWD yielded optimum response characteristics to a modulating rate of 200 kilobits per second when the pulses had a width of 0.2 microseconds and an amplitude of 7 volts.

An RF shaping network was used to convert the incoming data stream into RF pulses for excitation of the transmitter ASWD's. See Figure 5.

The circuit employs a free-running crystal controlled oscillator whose output of 21.35 megahertz is fed to one input of a multiplier circuit. The other input to the multiplier is driven by the information data stream. This process has the effect of gating the RF energy "on" during the occurrence of a data pulse and gating it off when no pulse is present. See Figure 6.

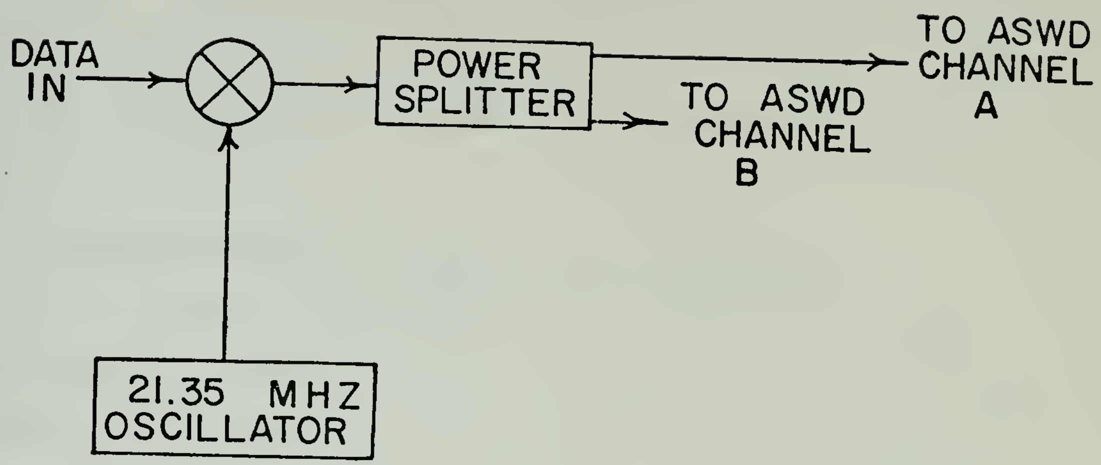

Figure 5. RF Pulse Shaping Network.

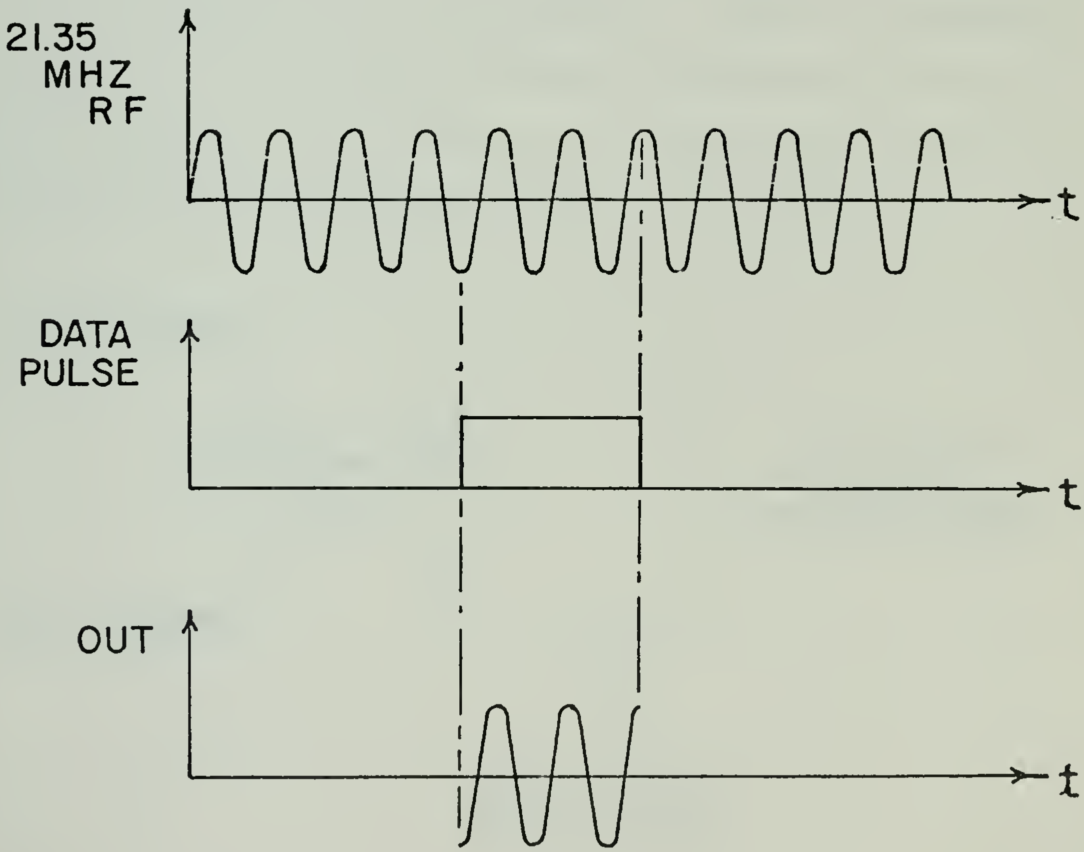

Figure 6. Voltage Waveforms of the RF Pulse Shaping Network.

(Notice, in Figure 6, that the pulse width is controlled by the width of the input data pulses.) The pulsed RF signal is fed to a power splitter to provide two identical outputs, which are necessary for driving the complementary coded ASWD's simultaneously.

2. Multiplexing the Complementary Waveforms

The output of the ASWD's after proper excitation consisted of two spread spectrum signals each with a center frequency of approximately 21.4 megahertz and a bandwidth of approximately four megahertz. Multiplexing of the two channels was performed by translating one of the channels to a center frequency of 30 megahertz. This 8.6 megahertz separation of the channels afforded a guard band that permitted adequate channel separation without the necessity of complicated space consuming filters. See Figure 7.

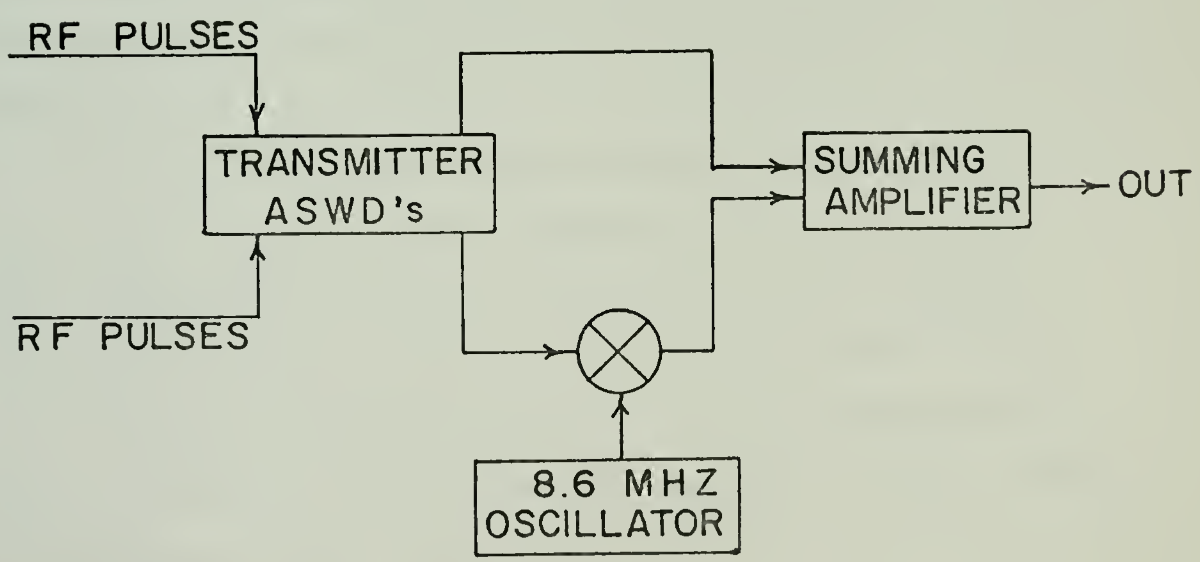

Figure 7. Frequency Multiplexer.

The Channel B output of the transmitter ASWD was fed into one input of a Watkins-Johnson, low-noise, wideband mixer (WJ-MI). The other input was a crystal controlled oscillator with an output frequency of 8.6 megahertz. The output of the mixer, which represented the Channel B waveform from the ASWD at a center frequency of 30 megahertz, was combined with the Channel A waveform to form a total spread spectrum signal of approximately 12 megahertz with a center frequency of 25.7 megahertz. It was this signal that was amplified and transmitted.

3. Amplification and Transmission of the Multiplexed Signal

With the multiplexed spread spectrum waveform defined, the problem of transmitting the signal was reduced to that of providing amplifiers with the appropriate gain and bandwidth characteristics. The wide bandwidth of approximately ten megahertz posed two separate types of problems.

The first problem was that of providing devices in the signal path with linear characteristics across the band. This linearity was necessary in order to prevent undesirable waveform distortions and cross modulation of the frequency multiplexed signals. Since compensating networks represented an uneconomical use of space, this method of providing a flat response was avoided. The desired amplifier response was obtained by using two stages of preamplification, provided by amplifiers obtained from a commercial supplier, and by using a power transistor with a

high gain bandwidth product. Features of the commercially available amplifiers included small size (approximately 1.5 oz) 30 decibels of power gain, flat response from 0.5 megahertz to 100 megahertz, and a three decibel noise figure.

The second problem that resulted directly from the wide bandwidth of the transmitted signal was that of noise power content. This noise power is proportional to the signal bandwidth and is equivalent to -104 dbm for a bandwidth of ten megahertz. This high $k_t b$ noise power was especially significant when the size constraint of the transmitter was taken into account. This effect is illustrated in the following sample calculation of received signal-to-noise ratio of a wideband system.

Noise power	-104 dbm	(10 MHz)
Transmitted power:	30 dbm	(1 Watt)
Antenna gains	0 db	
100 mile path loss	-119 db	
Receiver noise figure	5 db	
Total noise power	- 99 dbm	
Total signal power	- 89 dbm	
Signal-to-Noise ratio	10 db	

The above figures show the effect of a large bandwidth on signal-to-noise ratio. A bandwidth of only one megahertz would have resulted in a signal-to-noise ratio of 20 db.

Further calculations revealed that a transmitter operating in the S-band would be required to radiate approximately one watt for satisfactory reception of the spread spectrum signal at ranges up to 100 miles. Although this system was designed to operate in the high frequency range,

for ease of testing, the transmitted power was designed to be between one and two watts.

4. Reception and Amplification of the Spread Spectrum Signal

The problem of receiving and amplifying the spread spectrum signal was much like that of amplifying the signal in the transmitter. The wide bandwidth caused the major emphasis of the receiver to be on signal linearity and low noise susceptibility. The input stages of the receiver were designed around a commercially available wideband amplifier with a noise figure of three decibels and a power gain of 30 db. This low noise input amplifier caused the overall noise figure of the input amplifiers to be approximately four decibels. Subsequent amplifiers in the signal path of the receiver were required as interface elements between the various receiver functional devices. These amplifiers are described in detail in the following section on system integration.

5. Demultiplexing and Demodulating the Received Signal

The two channels of the spread spectrum were separated by active filters in the receiver signal path. See Figure 8.

The filter of Figure 8 labeled filter A was designed and built to have four megahertz bandwidth with a center frequency of 30 megahertz. The output of this filter was mixed with a local oscillator frequency of 8.6 megahertz to provide a difference frequency of 21.4 megahertz and a bandwidth of four megahertz. The filter of Figure 8 labeled

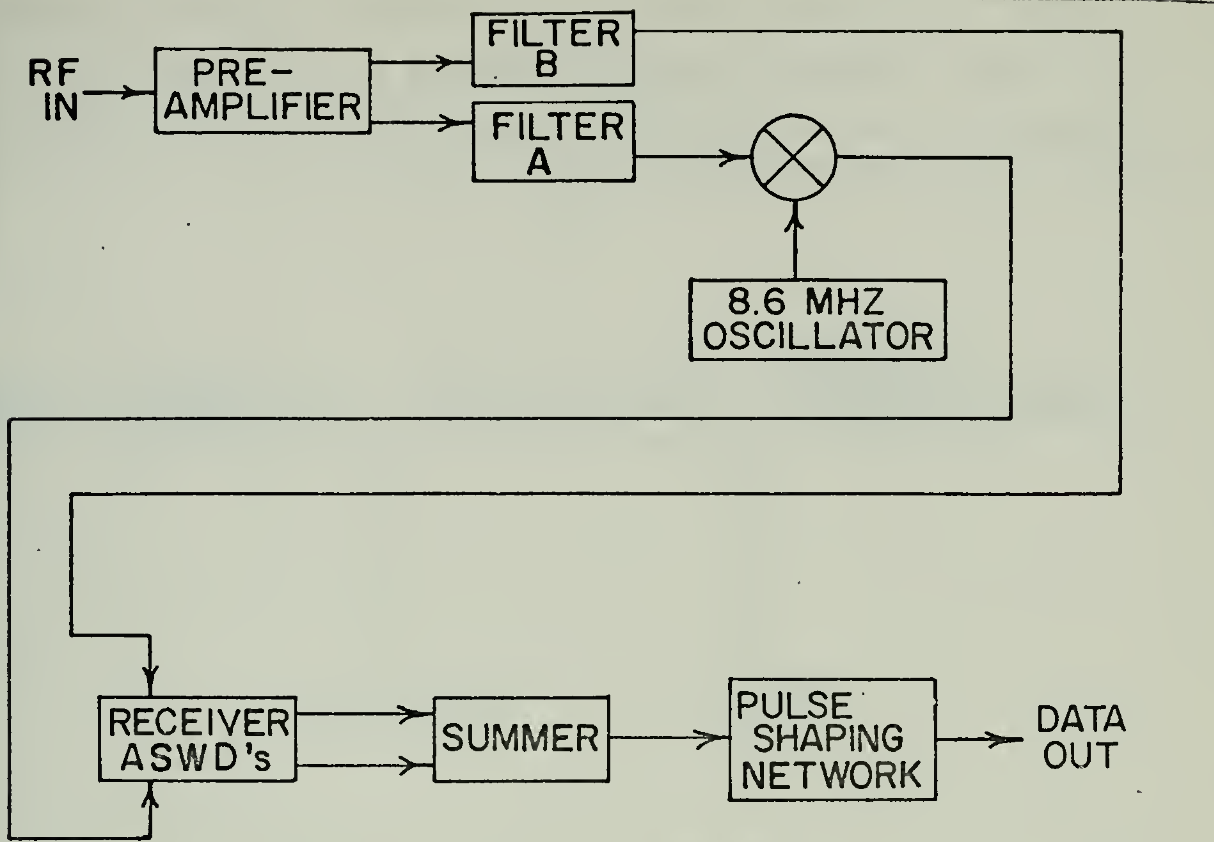

Figure 8. Block Diagram of Demultiplexing/Demodulation Scheme.

filter B was built to have a bandwidth of four megahertz centered about 21.4 megahertz. The signals from the two 21.4 megahertz channels were amplified and used to excite a pair of Golay complementary coded ASWD's identical to those of the transmitter section. These ASWD's acted as passive correlators for the signals of the two spread spectrum channels. See Appendix C. The outputs of the ASWD pair were fed to a summing amplifier to provide a single signal pulse each time a binary "1" was used to modulate the RPV transmitter. The summing amplifier is described in the following section.

A pulse shaping network was necessary to provide a suitable data train for the video processing equipment. See

Figures 9 and 10. Figure 9 is a functional block diagram of the pulse shaping network. Figure 10 gives a representation of the various waveforms from the blocks of Figure 9.

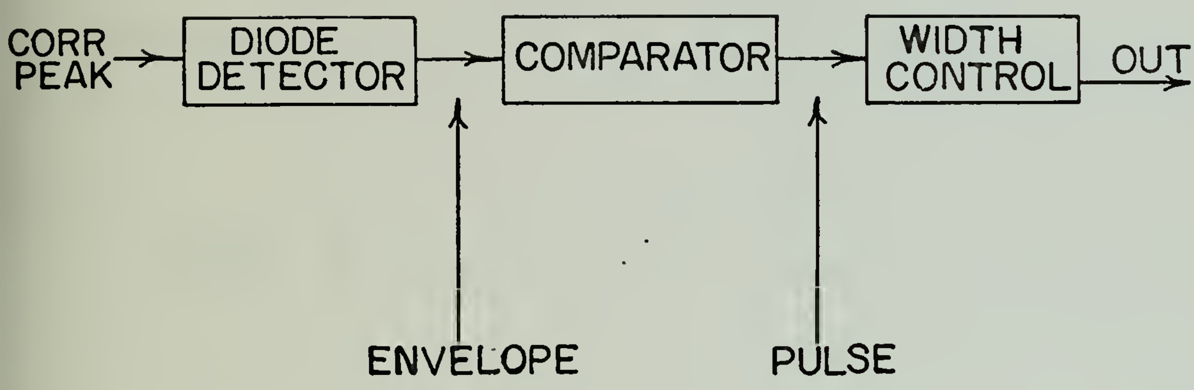

Figure 9. Block Diagram of the Pulse Shaping Network.

The correlation peaks from the summing amplifier were fed through a diode detector which extracted the envelopes from the signal pulses. This waveform was then fed to a variable threshold comparator which produced a pulse fixed width and amplitude whenever the input crossed a reference threshold level. The resultant pulse was used to trigger a monostable multivibrator whose output signal was compatible with video signaling equipment. A pulse width control allowed for slight instabilities in the timing synchronization hardware.

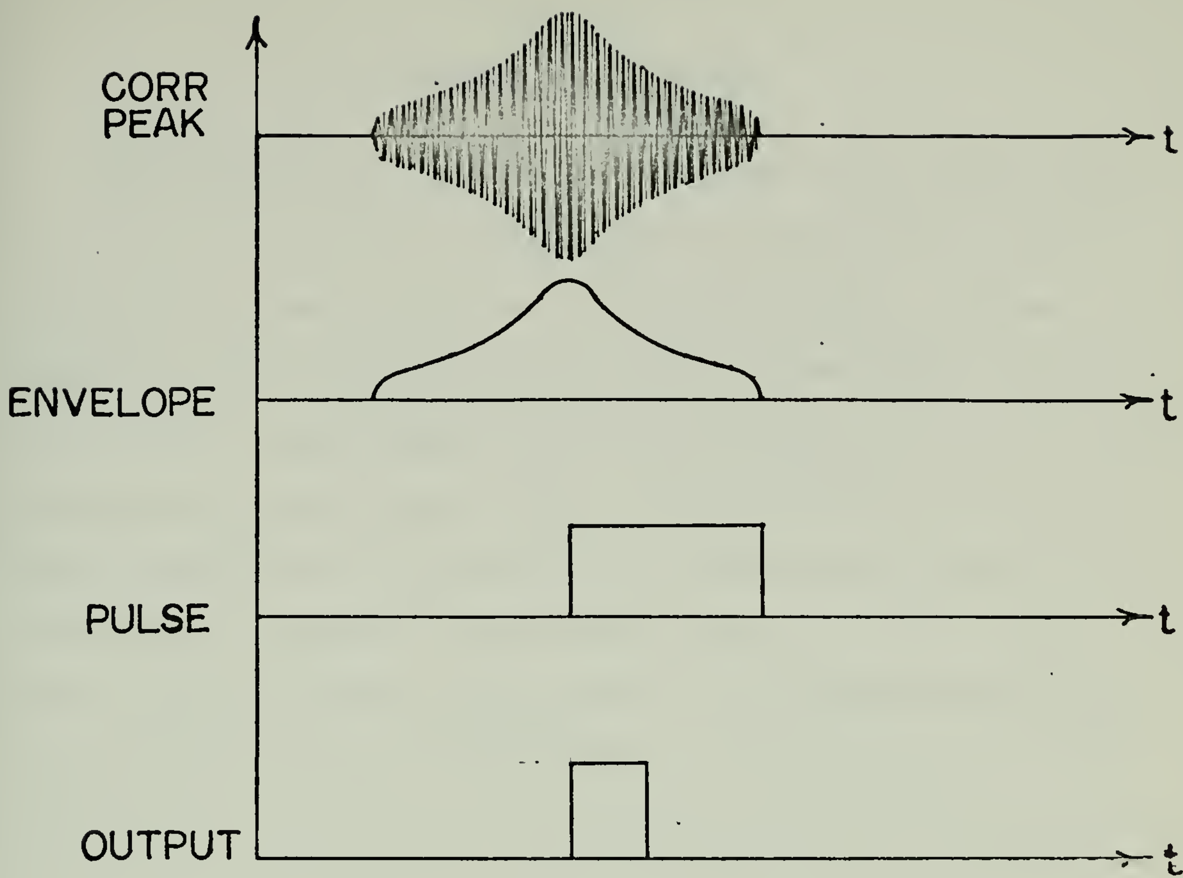

Figure 10. Representative Waveforms from Pulse Shaping Network.

IV. SYSTEM INTEGRATION

The circuits and devices described in the previous section required certain interfaces for consolidation of the independent elements into an integrated system. In many instances these interfaces consisted of simple BNC or SMA connectors. Other simple interfaces included direct wiring and coupling capacitors. But in most interface situations, buffer amplifiers, pre-amplifiers, summing amplifiers, or filters are necessary to make the output of a device compatible with the input requirements of the following device. This section describes the various interfaces used to supplement the circuits and devices described in the previous section. See Figures 11 and 12 for functional block diagrams of the transmitter and receiver sections of the completed system.

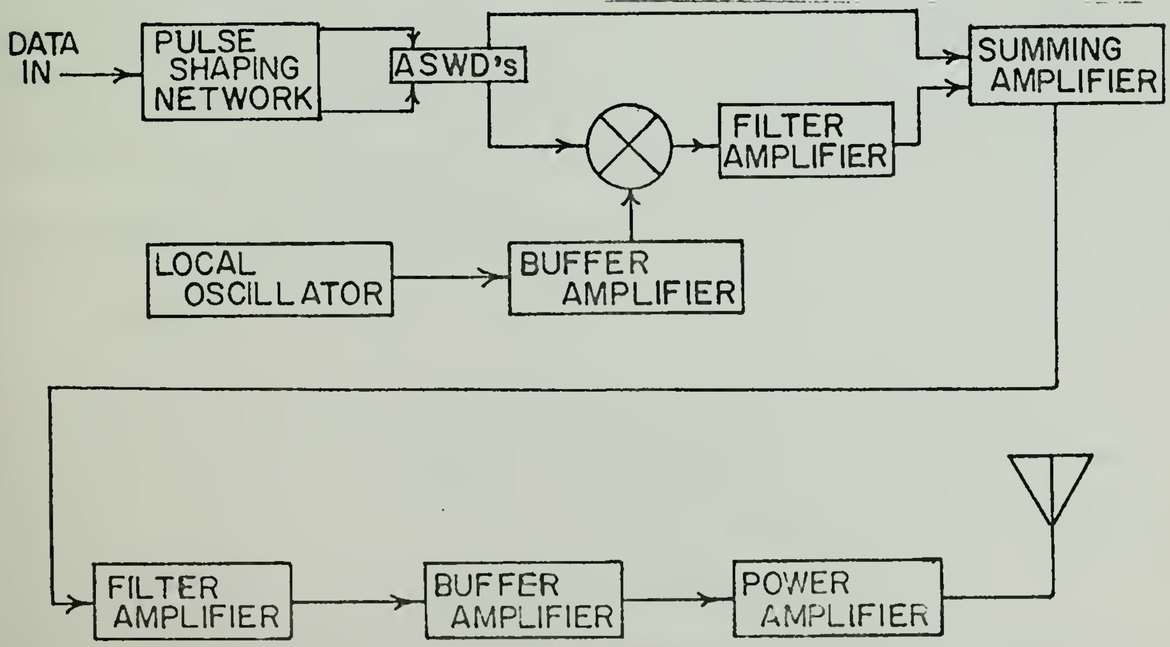

Figure 11. Transmitter Subsystem Functional Block Diagram.

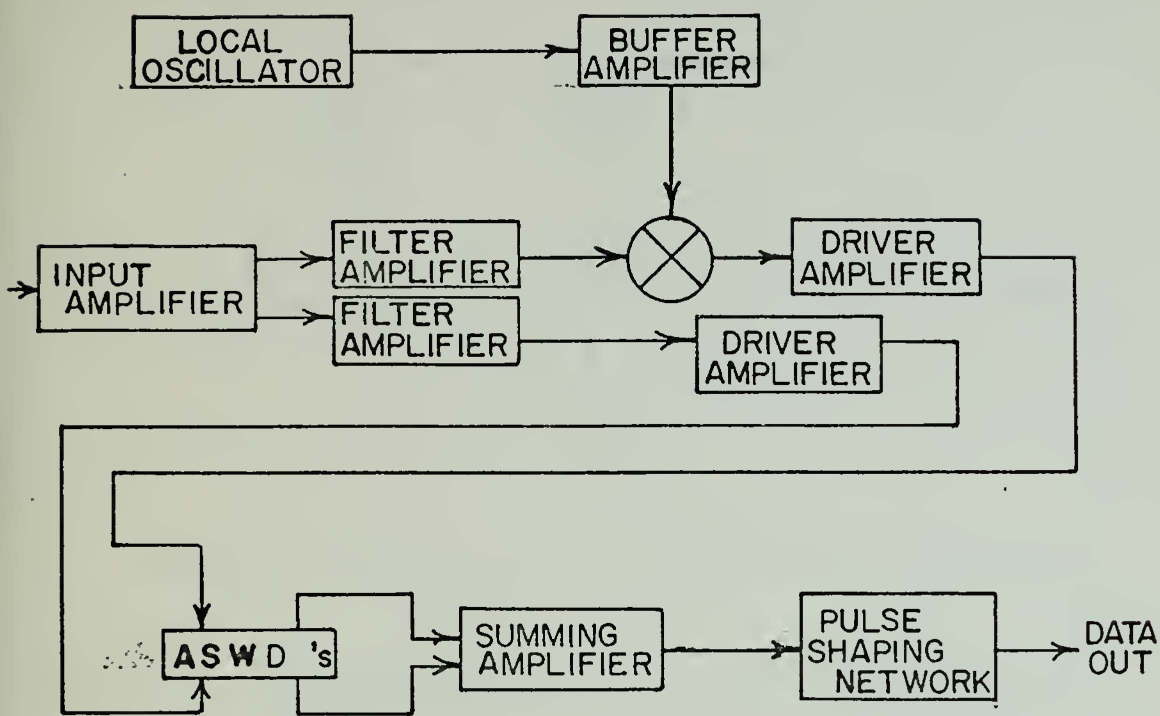

Figure 12. Receiver Subsystem Functional Block Diagram.

The blocks labeled "filter amplifier," "buffer amplifier," "driver amplifier," and "summing amplifier" are the interfaces that will be described in this section. Only those circuits that represent significantly different design approaches will be described in detail. Coupling capacitors and simple connectors will not be covered.

A. FILTERS

With the exception of simple RC filters and tuned LC circuits that are included in the devices described in the previous section, all of the filters designed and constructed for use in the system are analog active filters. The filters are represented by those blocks in Figures 11 and 12 which are labeled filter amplifiers.

See Figure 13 for a circuit diagram of the scheme used in the design of the active filters.

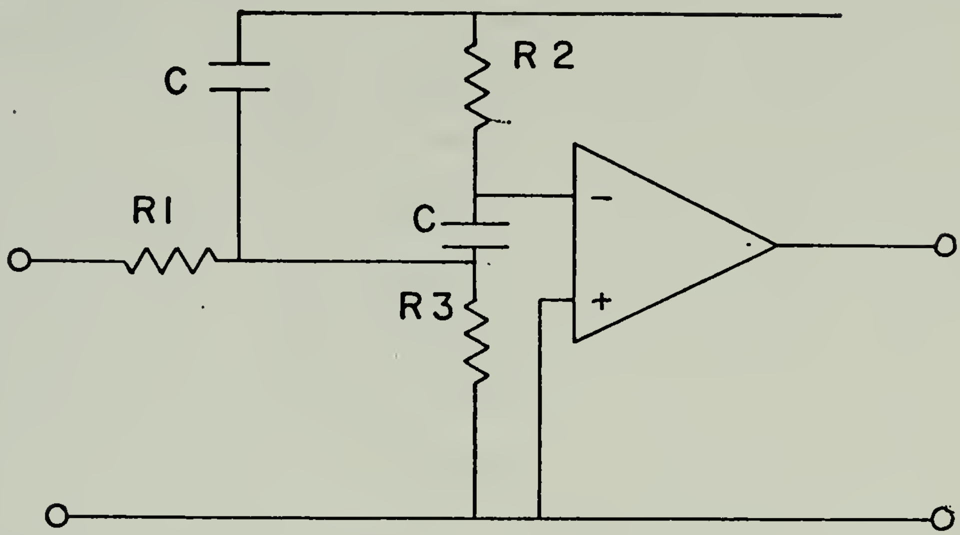

Figure 13. Circuit Diagram of a Typical Active Filter Amplifier.

Proper choices of the resistors labeled R1, R2, and R3 and the capacitors labeled C resulted in filters with appropriate center frequencies and bandwidths. The operational amplifiers were chosen to be capable of responding to frequencies in the desired range.

Let: Fo = the desired center frequency of the filter

BN = the desired bandwidth

G = nominal voltage gain

then, $R1 = 1 / 2\pi(BW)GC$

$R2 = 1 / \pi(BW)C = R1(2G)$

$R3 = (1 / 2\pi C[(2Fo^2/(BW)) - (BW)G])^{\frac{1}{2}}$.

Note that only the equation for R3 depends upon the center frequency. Therefore, the center frequency of the filter

can be adjusted without altering the bandwidth of the filter. Typical values for a representative filter used in the system follows:

$$Fo = 30\text{MHz}$$

$$BW = 4\text{MHz}$$

$$G = 10\text{db}$$

$$R1 = 408\Omega$$

$$R2 = 8160\Omega$$

$$R3 = 9.4\Omega$$

$$C = 10\text{pfd.}$$

B. AMPLIFIER INTERFACES

Amplifiers were required to make the outputs of some stages compatible with the inputs of the following stages. These interfaces are represented by those blocks of Figures 11 and 12 which are labeled "buffer amplifier," "summing amplifier," and "driver amplifier." The driver amplifiers were purchased from a commercial supplier. Features of these commercially available amplifiers include a three decibel noise figure, 30 decibels of power gain, BNC connectors, and small physical size (approximately 1.5 ounces). The other amplifiers were designed and built utilizing integrated circuit operational amplifiers with high gain bandwidth products and low noise characteristics. These amplifiers were designed using standard operational amplifier configurations. All of the IC chips used were readily available off-the-shelf items. Circular chips were used in preference to rectangular chips for circuit compactness.

Since both positive and negative voltages are readily available from RPV power supplies, amplifier interfaces were built to utilize dual polarity power in order to minimize the necessary external biasing circuitry. See Figure 14 for a representative circuit diagram of a typical buffer amplifier.

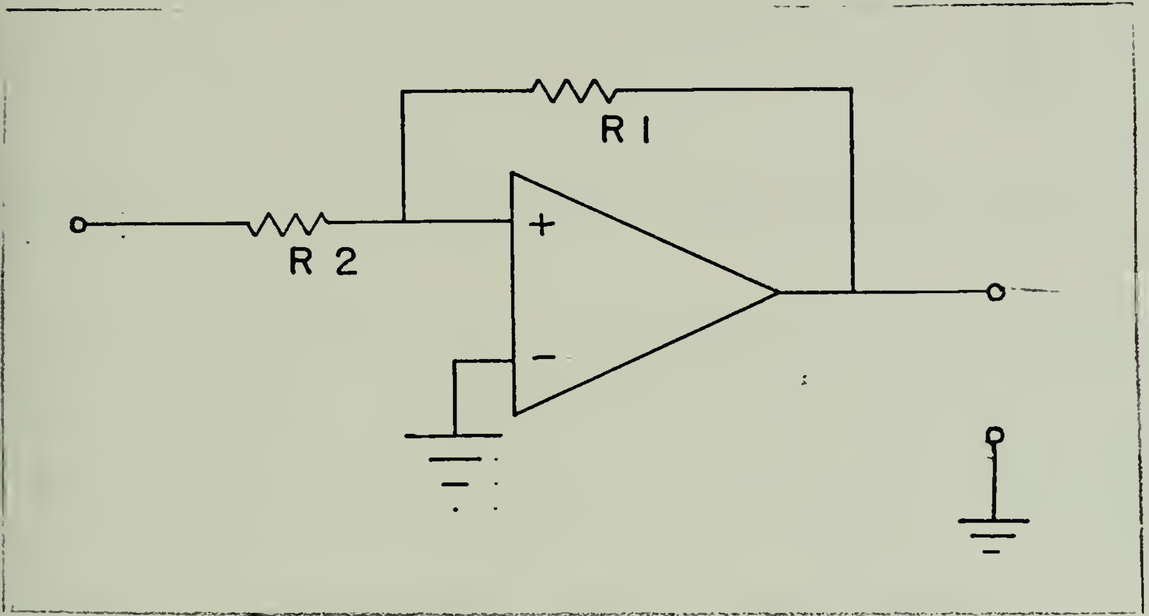

Figure 14. Circuit Diagram of a Typical OP-AMP Buffer Amplifier.

Resistors R2 and R1 are chosen to give the desired gain characteristics of the amplifier circuit. The voltage gain, G, of the circuit is given by the formula:

$$G = R1/R2.$$

This amplifier arrangement permitted adjustments of the amplifier gain to make it suitable for the input signal amplitude and thereby avoiding non-linearities from amplifier limiting.

Figure 15 represents a typical circuit diagram of a summing amplifier interface.

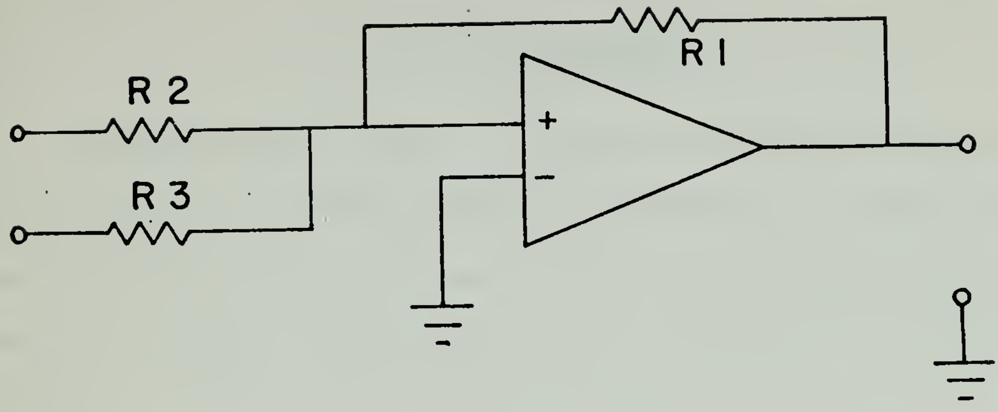

Figure 15. Circuit Diagram of a Typical OP-AMP Summing Amplifier.

In the summing amplifiers, the input resistors, R2 and R3, were balanced with the feedback resistor, R1, to give the desired gain characteristics as described in the paragraph above. R2 and R3 were also balanced to compensate for differences in the amplitude of the two input signals.

V. TEST PROCEDURES

System and subsystem testings were routinely performed as components were designed and fabricated. This testing was necessary to determine circuit waveform parameters so that the proper interfaces could be provided between the elements of the system. Whereas most of this testing represented standard engineering practices, and therefore will not be described in detail, significantly different procedures had to be utilized for the testing of the equipment's anti-jam capabilities. This section describes tests made on the communication system to determine its performance in interfering noise environments. The system was tested in the presence of two types of noise interference that are often encountered in communications jamming -- narrow band interference (or continuous wave (CW)) and wideband noise interference.

A. TESTING CONSIDERATION

For an understanding of the effect of processing gain (PG) on a system's performance, consider a system that will permit satisfactory receiver output when the received waveform has a signal-to-noise ratio of ten decibels or greater but does not produce an acceptable output when the input signal-to-noise ratio is less than ten decibels. A realization of ten decibels of PG in this system would permit satisfactory operation at a signal-to-noise ratio of unity

(0 db), and a realization of 20 decibels of PG would give the system the capability of satisfactory performance when the signal-to-noise ratio was a negative ten decibels. It is this feature of being capable of recovering a signal that is well below the level of intentional and incidental noises that provides the motivation for achieving maximum processing gain. The level of processing gain provided by an efficient system can be predicted by dividing the spread spectrum bandwidth by the bandwidth of the information signal. Since it was expected that the addition of Golay complementary coding would increase the amount of processing gain, a major consideration of the testing procedure was the determination of the effect of this coding on the system's processing gain.

B. TEST SETUP FOR NARROW BAND NOISE REJECTION MEASUREMENTS

The purpose of this examination was to determine not only the CW interference noise levels that the system could survive but also the frequency of narrow band interference that proved most troublesome. This test required accurate measurements of the signal and noise levels at the receiver input. See Figure 16 for a functional block diagram of the CW noise test setup.

This test arrangement permitted direct monitoring of the input signal-to-noise level by a wideband RF power meter. The frequency of the CW signal generator could be varied manually through the band of the spread spectrum waveform for a determination of the effect of frequency positioning on the system's performance.

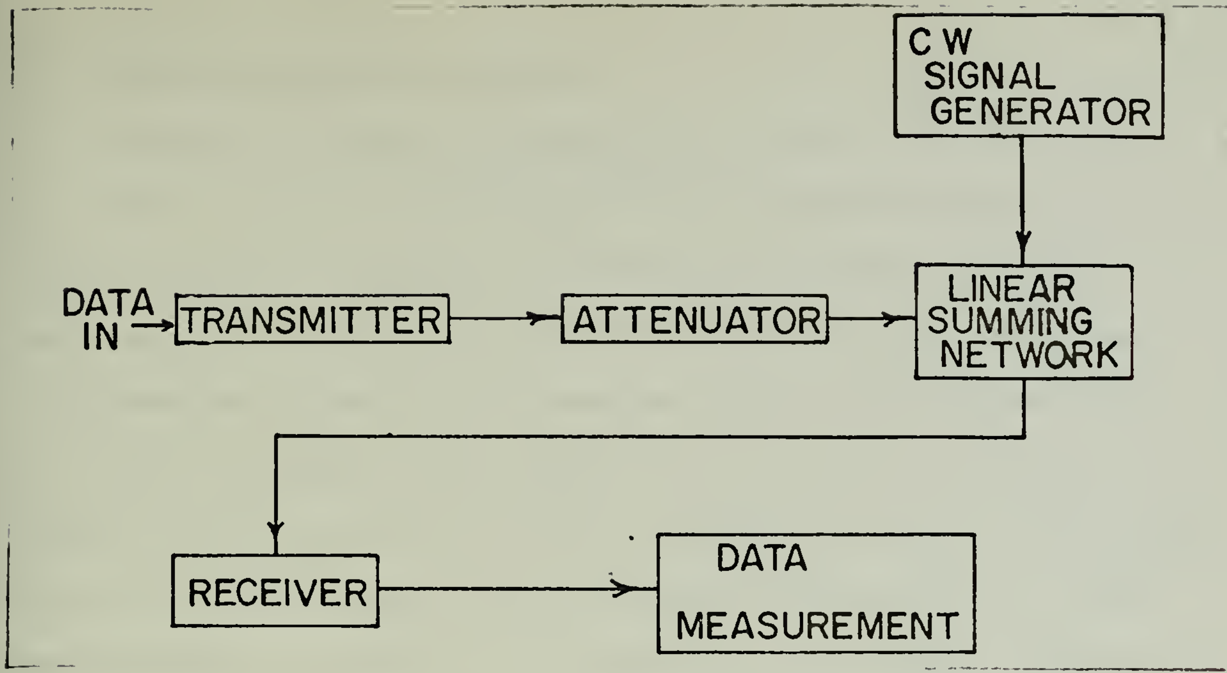

Figure 16. CW Noise Test Setup.

C. TEST SETUP FOR WIDEBAND INTERFERENCE REJECTION MEASUREMENTS

The purpose of this examination was to determine the susceptibility of the system to noise evely distributed throughout the band of the spread spectrum signal. This test gives a measure of the system's performance in the presence of wideband jamming, and it also gives a measure of the system's performance in the presence of natural noises which are not band limited. The test setup functional block diagram was the same as that of Figure 16 except the CW signal generator was replaced by a wideband noise generator. As in the previous interference measurement, the signal levels were measured directly. The noise power was determined by measuring the spectral power density of the noise source and computing the noise power in the spread spectrum band.

D. TEST EQUIPMENT AND DEVICES

Attempts were made to provide test conditions that were as realistic as could be realized in a laboratory test situation. However, an exact determination of the system's performance in various noise environment required accurate measurements of the test parameters. For this reason, signal and noise power measurements were made with a Hewlett Packard 232A wideband power meter instead of using a spectrum analyzer to determine relative signal-to-noise power ratios. Measurements were made at the inputs of the summing junction so that the resultant signal-to-noise ratios would be relatively independent of coxial transmission losses.

The CW noise interference signals were provided by an SG-85, URM-25D signal generator. Signal linearity and signal frequencies were verified by TEXTRONIX 422 oscilloscope measurements before injection into the receiver summing junction.

The wideband noise interference was provided by an ELGENCO Gaussian noise generator and by using a Watkins-Johnson Relcom mixer and a URM-25D signal generator for frequency translation.

Message quality measurements were made using a TSI USM 245A frequency counter. The transmitted data count was compared with the received data count as described in the following section to give a measure of transmission reliability in the presence of interfering noises.

VI. SYSTEM PERFORMANCE

This section reports on the results of the tests described in the previous section. Tests were conducted on the system to determine the effects of CW and wideband noise interference on system reliability. These test results give a measure of the anti-jam capability of the system as well as the ability of system to survive interference from natural and incidental noises.

The test results are tabulated in a manner to show the relationship between interference noise power levels and system reliability. Separate tables show the effect of CW interference frequency positioning. Each table is labeled according to the condition that prevailed during the test.

The first column of each table gives the total spread spectrum signal power delivered to the receiver input. Column two gives the total noise power delivered to the receiver input that is within the passband of the receiver. Columns one and two are expressed in dbm. Column three gives the signal-to-noise ratio at the receiver input, expressed in db. Column four gives the percentage of the messages sent that were actually received during a particular sample interval. The sample interval chosen for the test was one second.

During testing for transmission reliability, two types of error can be expected to occur:

FALSE ALARMS-----A message is received, but no valid message was sent.

MISSED MESSAGE---A valid message was sent, but no message was received.

The tests conducted on this system were designed so that the probability that a false alarm type error was small enough to be neglected. This condition was assured by adjusting interfering noise levels to be sufficiently low that the probability of a noise spike would exceed the receiver threshold level was negligible. So only missed messages had to be considered in a determination of system reliability. Messages were transmitted at data rates as specified on the individual tables, and a count was made of the valid messages actually received. The ratio of received messages to transmitted messages times one hundred gave the transmission reliability percentages listed in columns four of the tables.

A. SYSTEM PERFORMANCE IN VARIOUS CW JAM ENVIRONMENTS

Continuous wave interference signals were injected into the spread spectrum signal path at interference frequencies of 18 MHz, 19 MHz, 20 MHz, 21 MHz, 22 MHz, 23 MHz, 24 MHz, and 25 MHz. At each frequency the signal-to-noise ratio was decreased from a level at which the jamming signal had negligible effect on the system's performance to a level that caused the system to drop out of synchronization. CW jamming at frequencies of 18 MHz and below produced no significant system degradation so no data is included for jamming frequencies below 19 MHz. Also, waveform symmetry

about the 25 MHz point permits omission of test results at frequencies above 25 MHz.

Tables 1 through 7 gives a measure of the system's performance in the presence of CW interference for the conditions as specified by the tables' titles.

B. SYSTEM PERFORMANCE IN A WIDEBAND NOISE ENVIRONMENT

The system was subjected to a 5 MHz bandwidth gaussian noise interference to determine system performance in the presence of a wideband noise jam signal. A signal generator and a wideband mixer were used to position the gaussian noise waveform in the frequency band of 20 MHz to 25 MHz. This frequency positioning of the noise signal caused the noise to overlap the frequency band of only one of the spread spectrum signal channels. Equipment limitations precluded simultaneous jamming of the entire spread spectrum band with a wideband noise signal. The symmetry of the spread spectrum signal waveform about 25 MHz contributed to making this partial jamming of the signal bandwidth a meaningful test of the system's performance in the presence of wideband noise interference.

Table 8 gives a measure of the system's performance in the presence of wideband interference. The system did not drop synchronization within the range of noise power applied.

C. DISCUSSION OF SYSTEM PERFORMANCE

Tables 1 through 7 show the effect of CW noise jamming signals on the ASWD spread spectrum system. As predicted by theory, the jamming signal is most effective when the

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	500 KBITS
-10	-10	0	100	100
-10	- 9	- 1	100	100
-10	- 8	- 2	100	100
-10	- 7	- 3	100	100
-10	- 6	- 4	100	100
-10	- 5	- 5	100	100
-10	- 4	- 6	100	100
-10	- 3	- 7	100	100
-10	- 2	- 8	100	100
-10	- 1	- 9	100	100
-10	0	-10	100	100
-10	1	-11	100	100
-10	2	-12	100	100
-10	3	-13	100	100
-10	4	-14	100	100
-10	5	-15	100	100
-10	6	-16	100	99
-10	7	-17	100	99
-10	8	-18	99	99
-10	9	-19	99	99
-10	10	-20	99	

TABLE 1. SYSTEM PERFORMANCE IN THE PRESENCE OF A 19MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-10	-15	5	100	100	100
-10	-14	4	100	100	100
-10	-13	3	100	100	100
-10	-12	2	100	100	100
-10	-11	1	100	100	100
-10	-10	0	100	100	100
-10	- 9	- 1	100	100	100
-10	- 8	- 2	100	100	97
-10	- 7	- 3	100	97	95
-10	- 6	- 4	98	94	92
-10	- 5	- 5	95	91	91
-10	- 4	- 6	92	86	88
-10	- 3	- 7	90	79	84
-10	- 2	- 8	87	76	81
-10	- 1	- 9	85	73	80
-10	0	-10	84	DROP	77
-10	1	-11	80		76
-10	2	-12	77		74
-10	3	-13	73		DROP
-10	4	-14	71		
-10	5	-15	DROP		

TABLE 2. SYSTEM PERFORMANCE IN THE PRESENCE OF A 20MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-17.2	-27	10	100	100	100
-17.2	-26	9	100	100	100
-17.2	-25	8	100	100	100
-17.2	-24	7	100	100	100
-17.2	-23	6	100	100	100
-17.2	-22	5	100	100	100
-17.2	-21	4	100	100	100
-17.2	-20	3	100	100	100
-17.2	-19	2	100	100	100
-17.2	-18	1	100	100	100
-17.2	-17	0	100	99	100
-17.2	-16	- 1	100	99	99
-17.2	-15	- 2	98	97	99
-17.2	-14	- 3	94	94	97
-17.2	-13	- 4	91	90	93
-17.2	-12	- 5	86	85	87
-17.2	-11	- 6	80	81	82
-17.2	-10	- 7	76	75	77
-17.2	- 9	- 8	70	72	72
-17.2	- 8	- 9	DROP	DROP	DROP

TABLE 3. SYSTEM PERFORMANCE IN THE PRESENCE OF A 21MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR,	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-10	-14	4	100	100	100
-10	-13	3	100	100	100
-10	-12	2	100	100	100
-10	-11	1	100	100	100
-10	-10	0	100	100	100
-10	- 9	- 1	100	100	100
-10	- 8	- 2	100	100	100
-10	- 7	- 3	100	100	100
-10	- 6	- 4	99	100	99
-10	- 5	- 5	99	100	99
-10	- 4	- 6	99	100	99
-10	- 3	- 7	99	100	99
-10	- 2	- 8	97	99	98
-10	- 1	- 9	95	99	98
-10	0	-10	91	94	95
-10	1	-11	87	89	89
-10	2	-12	82	84	80
-10	3	-13	78	79	76
-10	4	-14	78	73	71
-10	5	-15	76	71	DROP
-10	6	-16	DROP	DROP	

TABLE 4. SYSTEM PERFORMANCE IN THE PRESENCE OF A 22MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-10	-10	0	100	100	100
-10	- 9	- 1	100	100	100
-10	- 8	- 2	100	100	100
-10	- 7	- 3	100	100	100
-10	- 6	- 4	100	100	100
-10	- 5	- 5	100	100	100
-10	- 4	- 6	100	100	100
-10	- 3	- 7	100	100	100
-10	- 2	- 8	100	100	100
-10	- 1	- 9	100	100	100
-10	0	-10	100	100	100
-10	1	-11	100	100	100
-10	2	-12	100	100	100
-10	3	-13	100	100	96
-10	4	-14	100	98	89
-10	5	-15	100	97	78
-10	6	-16	99	91	73
-10	7	-17	96	84	DROP
-10	8	-18	88	79	
-10	9	-19	81	73	
-10	10	-20	75	DROP	
-10	11	-21	DROP		

TABLE 5. SYSTEM PERFORMANCE IN THE PRESENCE OF A 23MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-10	-10	0	100	100	100
-10	- 9	- 1	100	100	100
-10	- 8	- 2	100	100	100
-10	- 7	- 3	100	100	100
-10	- 6	- 4	100	100	100
-10	- 5	- 5	100	100	100
-10	- 4	- 6	100	100	100
-10	- 3	- 7	100	100	100
-10	- 2	- 8	100	100	100
-10	- 1	- 9	100	100	100
-10	0	-10	100	100	100
-10	1	-11	100	99	100
-10	2	-12	100	99	100
-10	3	-13	100	99	95
-10	4	-14	100	91	91
-10	5	-15	98	87	84
-10	6	-16	95	82	77
-10	7	-17	90	78	72
-10	8	-18	82	73	DROP
-10	9	-19	75	DROP	
-10	10	-20	DROP		

TABLE 6. SYSTEM PERFORMANCE IN THE PRESENCE OF A 24MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-10	-10	0	100	100	100
-10	- 9	- 1	100	100	100
-10	- 8	- 2	100	100	100
-10	- 7	- 3	100	100	100
-10	- 6	- 4	100	100	100
-10	- 5	- 5	100	100	100
-10	- 4	- 6	100	100	100
-10	- 3	- 7	100	100	100
-10	- 2	- 8	100	100	100
-10	- 1	- 9	100	100	100
-10	0	-10	100	100	100
-10	1	-11	100	100	100
-10	2	-12	100	100	100
-10	3	-13	100	100	100
-10	4	-14	100	100	100
-10	5	-15	100	100	100
-10	6	-16	100	100	100
-10	7	-17	100	100	100
-10	8	-18	100	100	100
-10	9	-19	100	100	100
-10	10	-20	100	100	99
-10	11	-21	100	98	98

TABLE 7. SYSTEM PERFORMANCE IN THE PRESENCE OF A 25MHz CW JAM SIGNAL

SIGNAL POWER, dbm	NOISE POWER, dbm	SNR, db	QUALITY PERCENT 200 KBITS	300 KBITS	500 KBITS
-10	-15	5	100	100	100
-10	-14	4	100	100	100
-10	-13	3	100	100	100
-10	-12	2	100	100	100
-10	-11	1	100	100	100
-10	-10	0	100	100	100
-10	- 9	- 1	99	99	100
-10	- 8	- 2	96	99	100
-10	- 7	- 3	94	96	96
-10	- 6	- 4	90	91	91
-10	- 5	- 5	90	88	86
-10	- 4	- 6	88	86	85
-10	- 3	- 7	84	83	85
-10	- 2	- 8	83	81	85
-10	- 1	- 9	81	80	83
-10	0	-10	80	80	81
-10	1	-11	79	79	80
-10	2	-12	76	79	80

TABLE 8. SYSTEM PERFORMANCE IN THE PRESENCE OF A WIDEBAND JAM SIGNAL

frequency of the CW jammer coincides with the center frequencies of the spread spectrum channel waveforms. Jamming signals displaced from the center frequency of a spread spectrum channel waveform by 1 MHz required approximately 6 to 7 more db of power to have as much effect on the system's performance as a jamming signal in the center of a channel waveform. Each additional displacement by 1 MHz showed a reduction of about 5 db in the effectiveness of the jamming signal.

With a CW jam signal placed in the center of a channel waveform, the system copies solidly at a signal-to-noise ratio of -1 db and dropped synchronization at a signal-to-noise ratio of -9 db.

These tests also showed that the system can be used to transmit data at a data rate of 500 KBITS without more than a 2 db degradation in processing gain.

The wideband noise test revealed that the system is capable of copying solidly in a wideband noise environment which yields a signal-to-noise ratio of -1 db, and it will copy without losing synchronization at signal-to-noise ratios of -12 db or greater.

The synchronization referred to in the previous paragraphs has nothing to do with the transmitter-to-receive synchronization. No such synchronization is necessary in a simple ASWD system. The method used to record received message quality involved a frequency counter which would lose synchronization with the pulse train when noise power became sufficiently large to produce triggering spikes.

This was not considered to be an unrealistic constraint on the system, as any data train requires some method of synchronization before the data can be processed.

VII. CONCLUSION

This thesis effort resulted in a laboratory system capable of satisfying many of the objectives sought at the outset of the investigation. Based on the results of tests performed on the system, a processing gain of approximately 19 db was realized. Since the bandwidth spreading ratio of 20 to 1 predicts 13 db of processing gain. It is felt that 6 db of processing gain are attributable to the Golay coding scheme of the ASWD pair. Whereas the system has not yet been consolidated into a compact form, the transmitter when transformed from a laboratory system to a portable "toy" should weight between 16 and 24 ounces, and the receiver should weigh just over three pounds.

The specification of simplicity was completely satisfied as witnessed by the design and fabrication sections of this thesis. The fact that only a laboratory system was tested and reported on instead of a completed portable system is a factor imposed by time and the unpredictable shortages of critical devices and components.

Several areas of follow-on work are suggested by the efforts and results of this thesis work. Perhaps the most obvious one is improvements on the spread spectrum system capable of RPV usage at S-band frequencies. A longer sequence of ASWD codes could be used to give the system an improvement in both communications security as well as transmission security. A PN code modulator could be added

to the system to give each further communications security and processing gain.

Another area of follow-on work suggested by this thesis is a thorough investigation to determine the exact effect of Golay complementary coding on the processing gain of a spread spectrum system. Much of this investigation could be performed using simple MODEM's without the necessity of signal transmissions.

Covert transmission of video data is within the reach of current technology. This is the driving force that makes spread spectrum research challenging and rewarding.

APPENDIX A: ACOUSTIC SURFACE WAVE DEVICES AS SPREAD SPECTRUM MODULATORS

ACOUSTIC SURFACE WAVE DEVICE BACKGROUND

Acoustic surface wave devices operate on the principal that a physical wave will propagate on the surface of an elastic solid, a principal investigated by Lord Rayleigh in the nineteenth century. These waves have come to be known as "Rayleigh Waves." It was discovered that in the case of a piezoelectric solid, these waves can be excited electrically, and conversely can be reconverted to electrical energy. Figure 17 depicts a piezoelectric crystal, such as quartz, and serves to illustrate a method in which a surface wave may be induced on the crystal.

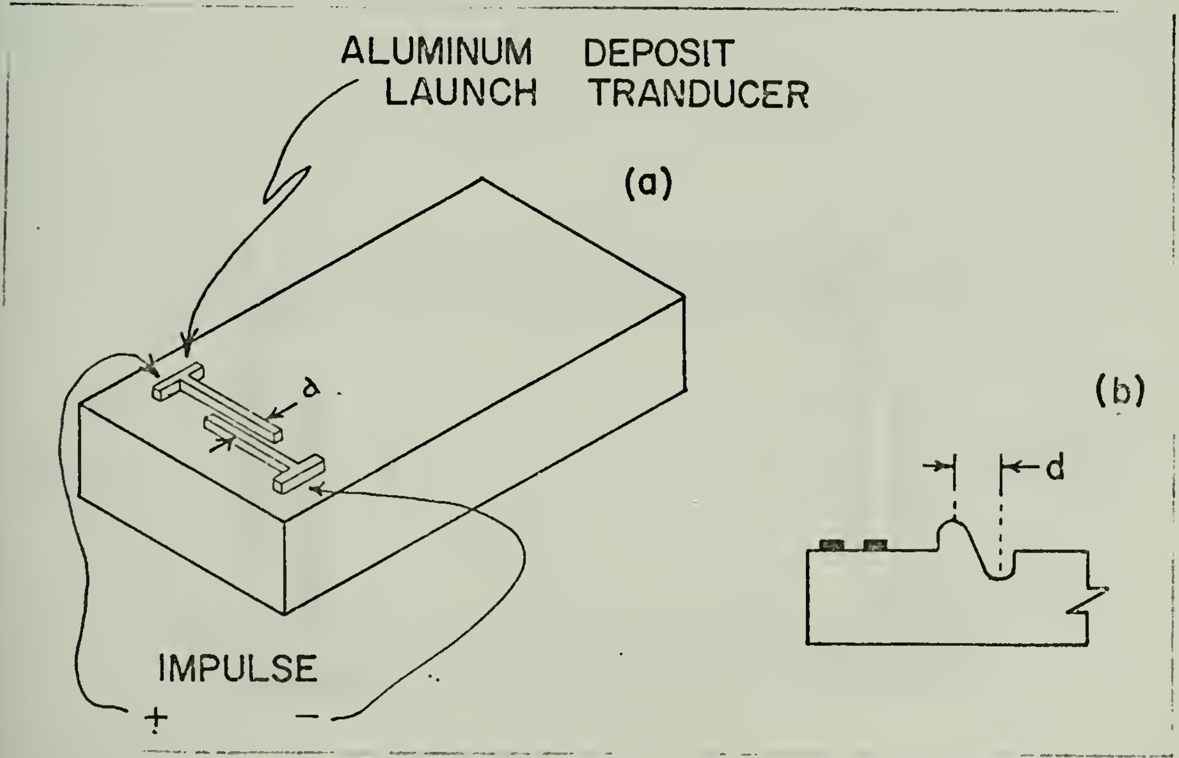

Figure 17. (a) Piezoelectric Crystal, with Transducer.
(b) Cross Section, Showing Launched Surface Wave.

When a narrow pulse is applied to the launch transducer, the piezoelectric nature of the crystal will transform the impulse response of the transducer (one sinusoidal cycle in this case) into a physical surface wave. A bulk wave will also be produced, but experiments have shown that these internal waves decay significantly after travelling a distance of one wavelength into the solid. Therefore, the majority of the energy is transformed into a wave travelling on the surface of the crystal. The propagation velocity of the surface wave is approximately 10^{-5} that of free space EM waves, thus the wavelength is similarly scaled. By carefully spacing the "fingers" of the transducers, the signal frequency may therefore be specified.

As an example, consider the crude device shown in Figure 18.

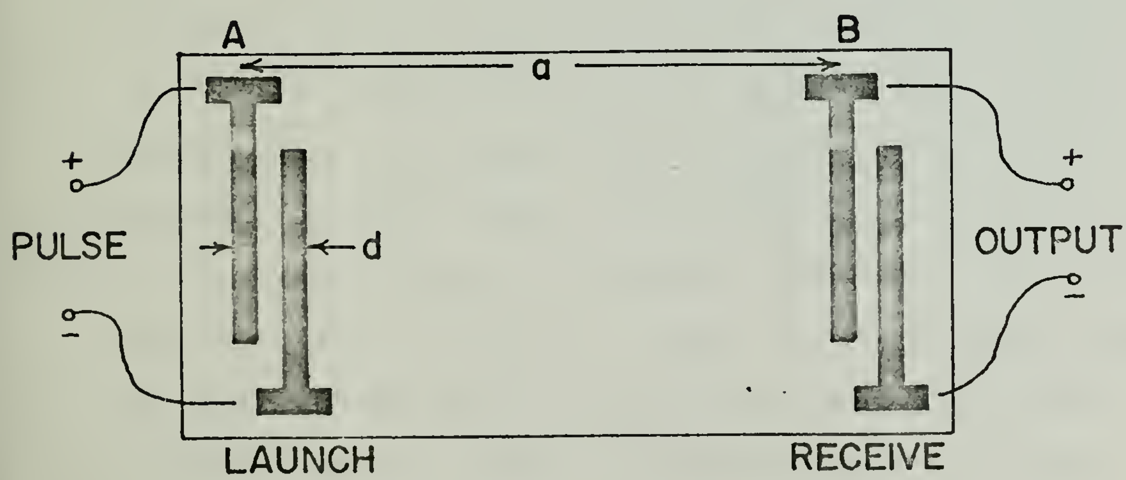

Figure 18. Surface Wave Device with Launch and Receive Transducers.

In this device, a receive transducer has been added to the right end which will covert the travelling surface wave to electrical energy by utilizing the piezoelectric crystal properties in the reverse manner as that of the launch transducer. Consider the dimensions to be as follows:

d: 0.03 m

a: 45.0 m.

The dimension "d" will be equal to ½ the surface wave wavelength, so the overall wavelength is 0.06 m. Using the 10^{-5} conversion factor, it can be seen that the free space equivalent wavelength is 6 meters, which corresponds to a frequency of 50 megahertz. Thus, when an impulse is applied to the launch transducer, one cycle of 50 megahertz RF will be launched on the crystal surface. This is due to the convolution of an impulse with the one finger-pair launch transducer.

Since the dimension "a" equals 45 mm, and the propagation velocity of the surface wave is 3×10^3 m/sec, then the propagation time to the receive transducer will be 15 microseconds. At that point, the surface wave will be coupled out by the receive transducer, and hence one cycle of 50 MHz RF will appear at the output. Notice that using the single finger pairs shown, one could feed pulse-modulated 50 MHz signals at the launch end of the device and effect a 15 m/sec delay of these pulses to the receive end. It is interesting to note that it would take 4500 meters of coaxial cable to realize this same delay.

Even with the crudest example, an obvious application of the surface wave device has become apparent: a simple delay line. Similar analysis can show how to adapt this simple model for use as a circulator as well.

In Figure 18, several receive transducers could have been placed along the propagation path, thus yielding what is known as a tapped delay line.

A few observations should be pointed out. In the simple models shown thus far, one finger pair in a transducer corresponds to one wave period. This property can be extended to any number of cycles desired. That is, a launch transducer with 13 finger pairs would launch 13 cycles of surface wave, and if this wave were coupled out by a receive transducer or n finger pairs, then (13 + n - 1) cycles of RF energy would be generated at the output terminals. The energy coupled out by the receive transducers can be phase coded by simple arrangement of the transducer fingers. Figure 19 illustrates how this is done.

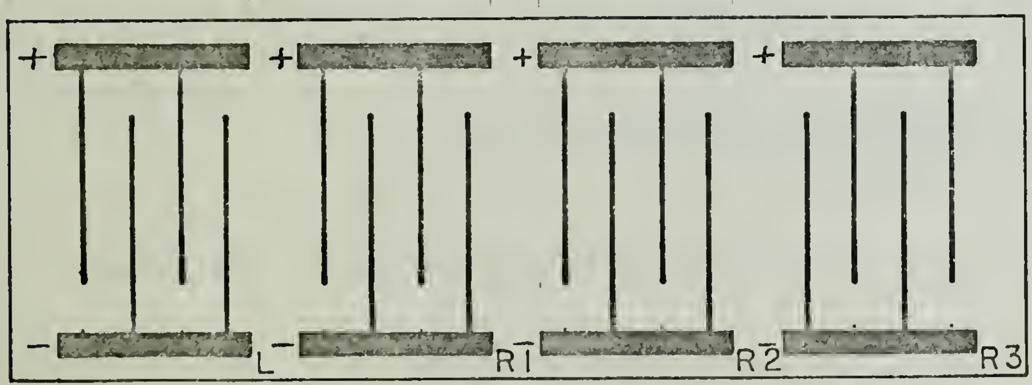

Figure 19. Phase Coding of the Output Transducers.

The receive transducers R1 and R2 have the same relative finger arrangement as the launch transducer, and thus will produce RF patterns "in phase" with the acoustic wave. Transducer R3, however, has a finger pattern opposite to that of the others and accordingly will produce a wave 180 degrees out of phase with the launched wave. Therefore, phase coded patterns can be physically implemented on the device.

The receive transducers may also be amplitude coded, thus giving rise to any wave shape the designer may desire as the device output. Figure 20 shows some simple examples

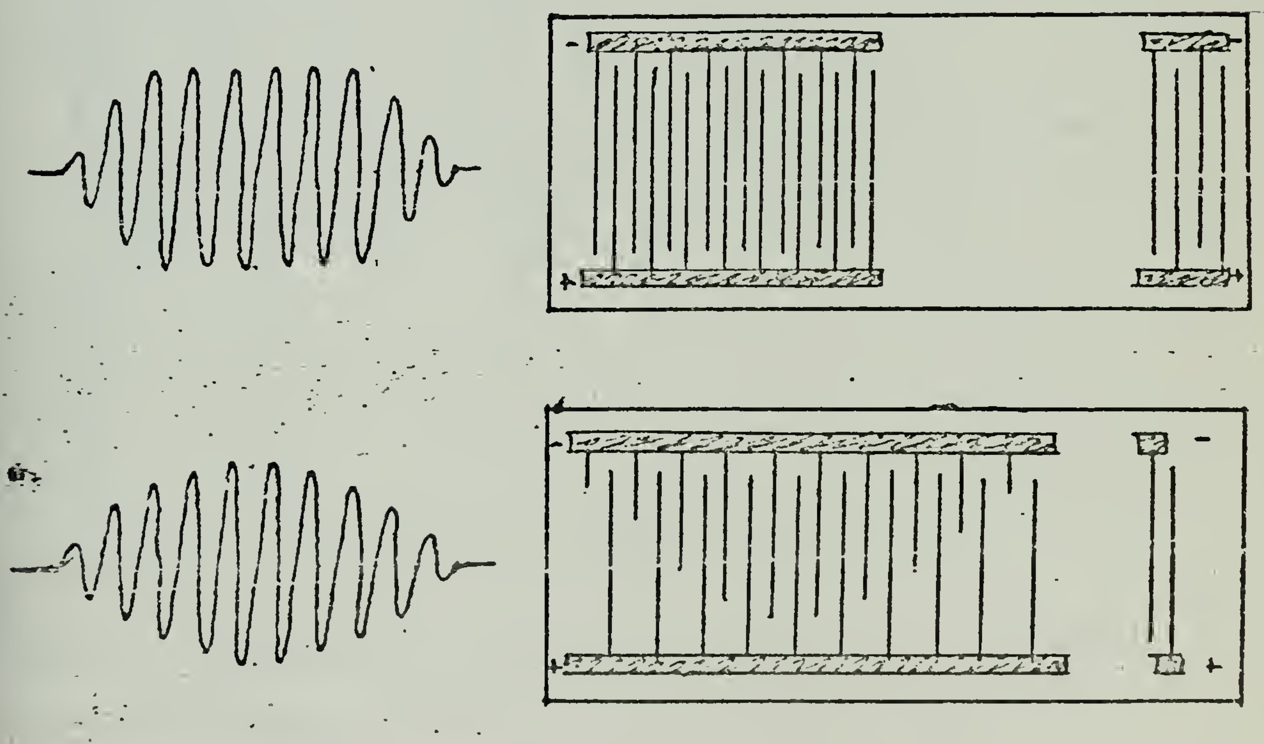

Figure 20. Amplitude Coded Waveforms.

Of interest is the fact that phase coding and amplitude coding can be accomplished simultaneously, but most current efforts tend to favor exploitation of the phase coding feature.

BI-PHASE CODING OF BINARY SEQUENCES

Consider the following four-bit binary sequence:

1 1 0 1.

This sequence can be represented by a bi-phase coded pulse train, assigning phase phi to a logical 1 and phase phi + 180 degrees to a logical 0. This train would appear as shown in Figure 21.

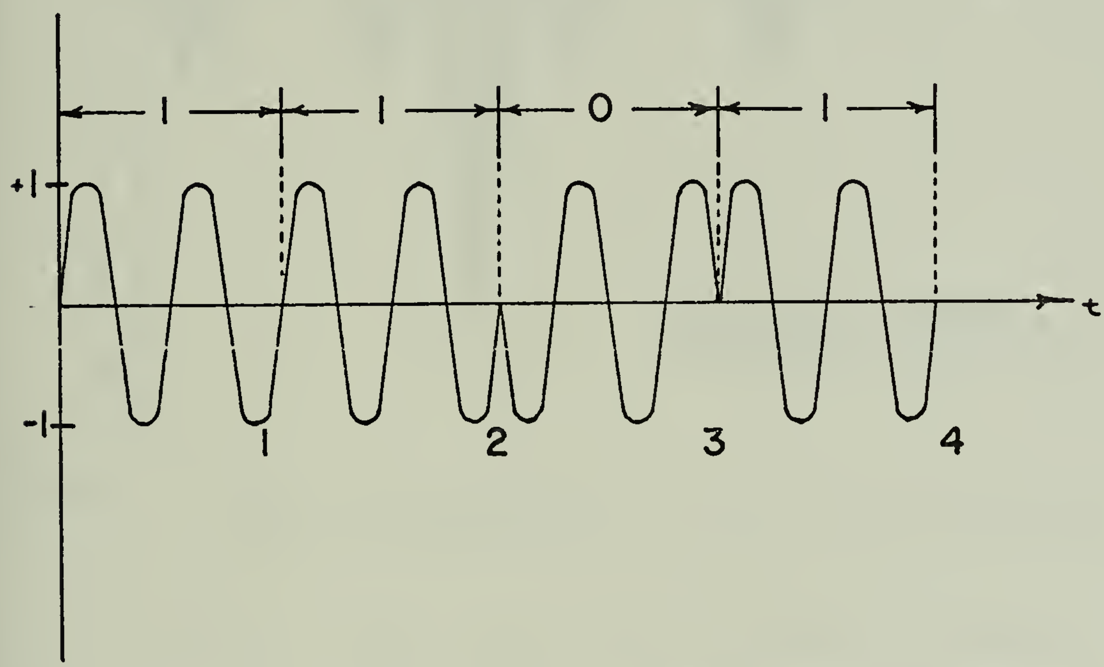

Figure 21. Bi-Phase Coded Binary Sequence 1 1 0 1.

If the time autocorrelation of this waveform were computed, the result would be a pulse train of length seven, with relative phases and amplitudes as shown in Figure 22.

Observe that a dominant center lobe is created (whose relative amplitude equals the length of the original sequence) along with sidelobes which are significantly below the center lobe level. This property is inherent in certain

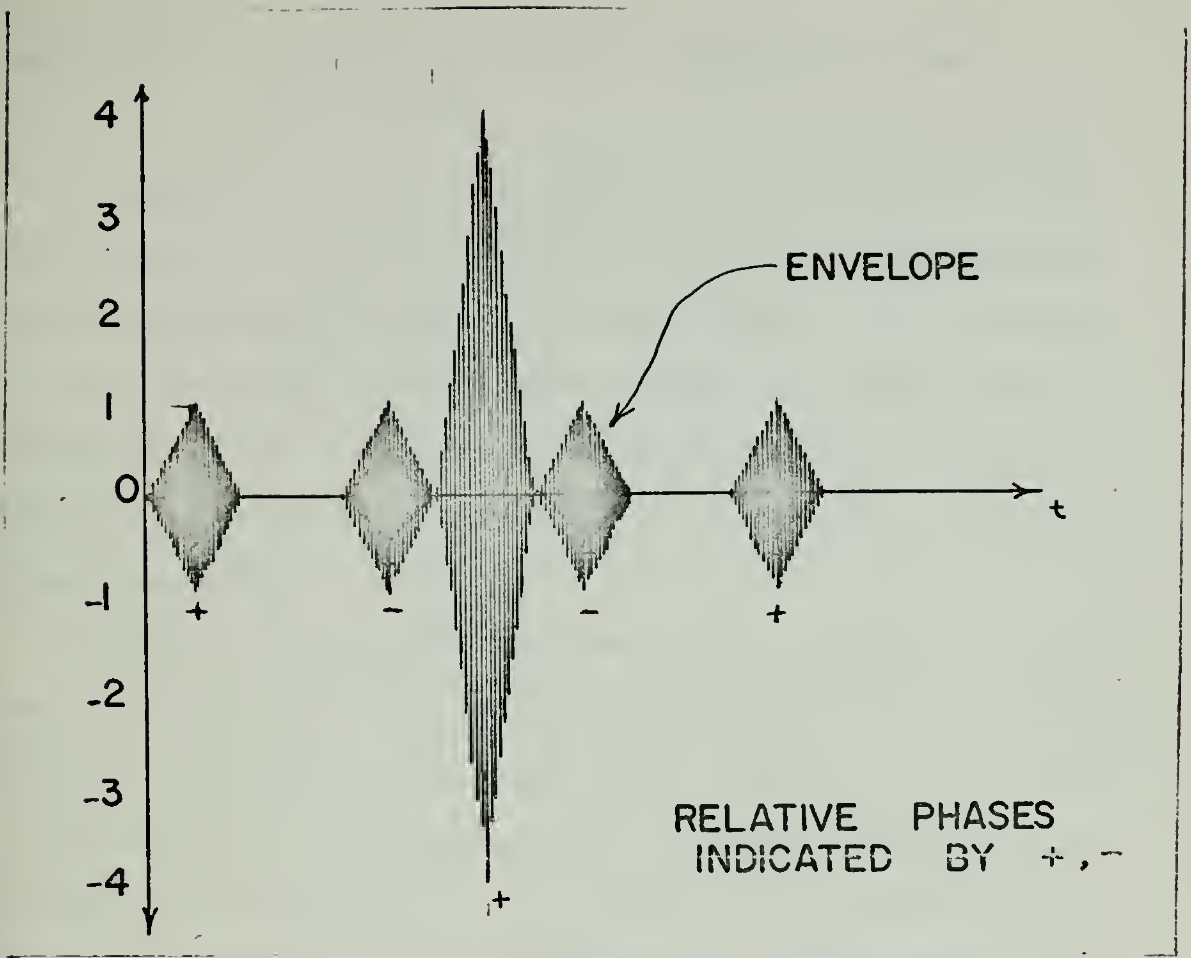

Figure 22. Time Autocorrelation Function of the Waveform Shown in Figure 21.

types of maximal length codes, and is crucial to spread spectrum forms of transmission. Notice that the sidelobes are symmetric both in amplitude and phase. This feature is of significance in certain coding schemes, particularly in complementary coded pairs utilizing Golay sequences.

APPENDIX B: GOLAY COMPLEMENTARY SEQUENCES IN ASWD APPLICATIONS

Golay complementary sequences are those binary codes which have an infinite correlation peak to peak ambiguity ratio when detected with a matched filter. See Reference 4 for a detailed theoretical treatment of these codes and Reference 3 for a discussion of applications. A simple example demonstrating this will be given here for illustrative purposes.

Suppose it is given that the binary sequences 1 1 1 0 and 1 1 0 1 are in fact Golay complements. These sequences could be represented by phase coded waveforms as discussed in Appendix A by assigning to two separate pulse trains, A and B, the relative + + + - and + + - +, respectively. The two sequences could be generated by two distinct surface wave devices. To generate the autocorrelation functions of these sequences, a minor modification to the devices must be made in the form of an added transducer as shown in Figure 23.

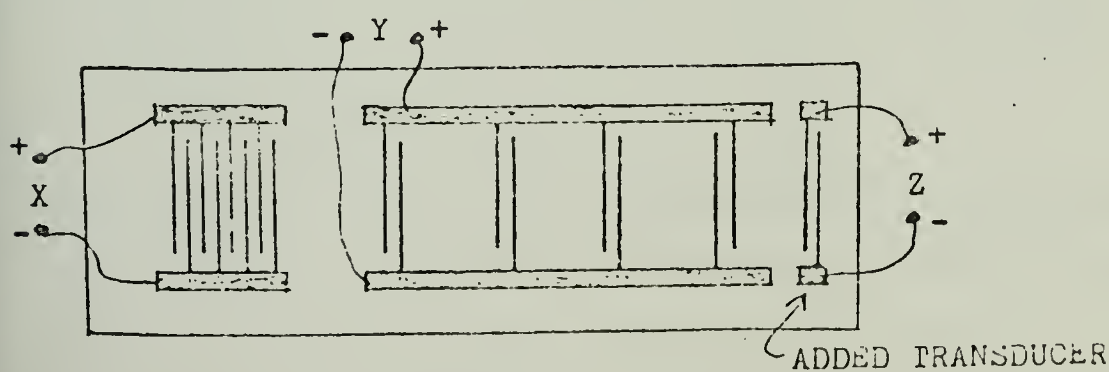

Figure 23. ASWD Modification for Correlation Process.

Assuming one device for each of the codes, the process would be as follows: If both ASWD's were excited by the same impulse, applied at port X, then the two waveforms coupled out from port Y would be of equal length and would be phase coherent. A matched filter technique would then be used to generate the autocorrelation functions. Consider the case of code A. If the bi-phase representation of this code was fed to port Y of its matched ASWD, then the autocorrelation function R_{AA} would appear at port Z. Tabular computation of R_{AA} is given in Table 9.

TIME	A*A	RAA
1	+++- +++-	- 1
2	+++- +++-	- 1 0
3	+++- +++-	- 1 0 1
4	+++- +++-	- 1 0 1 4
5	+++- +++-	- 1 0 1 4 1
6	+++- +++-	- 1 0 1 4 1 0
7	+++- +++-	- 1 0 1 4 1 0 - 1

Table 9. Computation of R_{AA}.

The sequence -1 0 1 4 1 0 -1 represents the relative phases and amplitudes of the pulses comprising the function R_{AA}. Similar analysis for code B would yield R_{BB} equal to 1 0 -1 4 -1 0 1. The property of the Golay codes is such that the algebraic sum of the autocorrelation functions

must be obtained to produce the resultant correlation function. This function is shown in Figure 24. The resultant correlation sequence is then 0 0 0 8 0 0 0, which exhibits the property of an infinite major peak to sidelobe

R_{AA}:	-1 0 1 4 1 0 -1
R_{BB}:	1 0 -1 4 -1 0 1
$R_{AA} + R_{BB}$:	0 0 0 8 0 0 0

Figure 24. Summing R_{AA} and R_{BB} to Obtain the Correlation.

peak ratio, in addition to a 3 db major peak gain over the single code case. Since this is the best theoretically obtainable correlation result (least probability of ambiguity between the major peak and sidelobe peaks,) then it was hypothesized that in practice it would also be the best result.

The increased processing gain (conventional codes would yield a final correlation peak of amplitude 4 with the addition of existing sidelobes) imply that in a noisy environment this coding scheme would be superior.

APPENDIX C: CHARACTERISTICS OF SYSTEM ACOUSTIC SURFACE WAVE DEVICES

Appendix A of Reference 2 contains a detailed description of the electrical characteristics and of the coding scheme ASWD's used in the transceiver system of this thesis. Relevant portions of the description are repeated here for report continuity purposes.

Figure 25 shows a photograph of one of the ASWD's as it was connected in the circuit. The casing in which the device is mounted was milled out of a solid block of aluminum

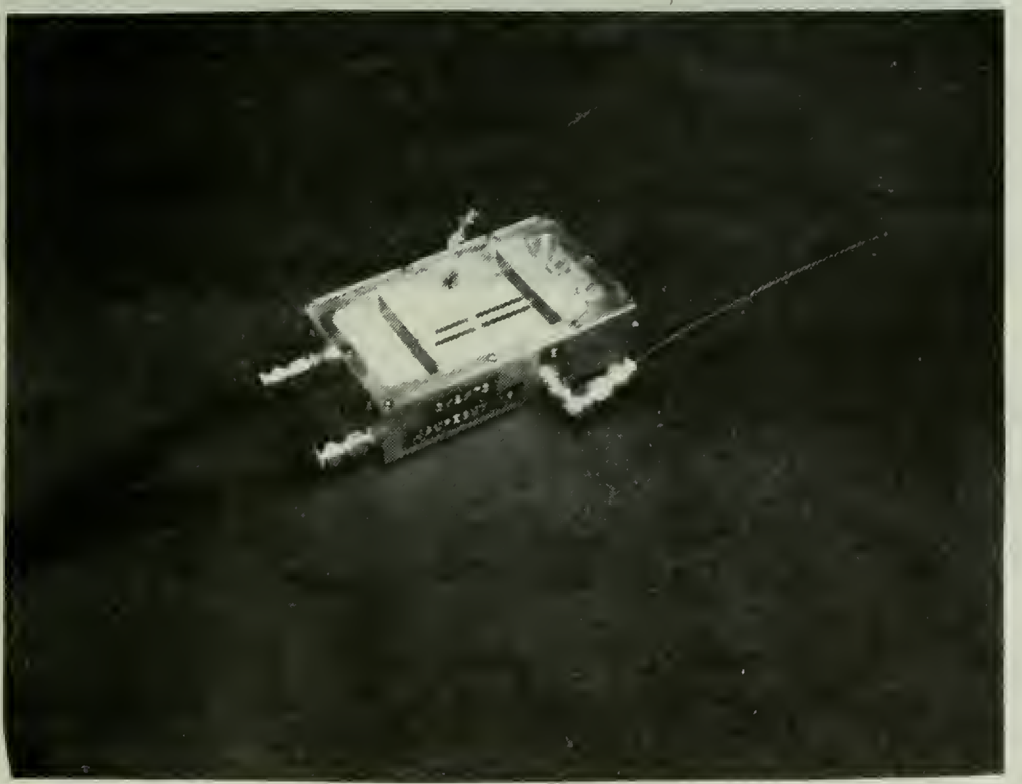

Figure 25. ASWD with External Connections.

that measures 4 inches long by 2.2 inches wide by one inch thick. The top of the casing, seen in the photograph, was milled to allow sufficient space to mount the quartz substrate

and to allow the inclusion of electrical feedthroughs from the external connection cavities on the reverse side of the casing, shown in Figure 26. There are six of these cavities, one for each of the three ports of each device, and each

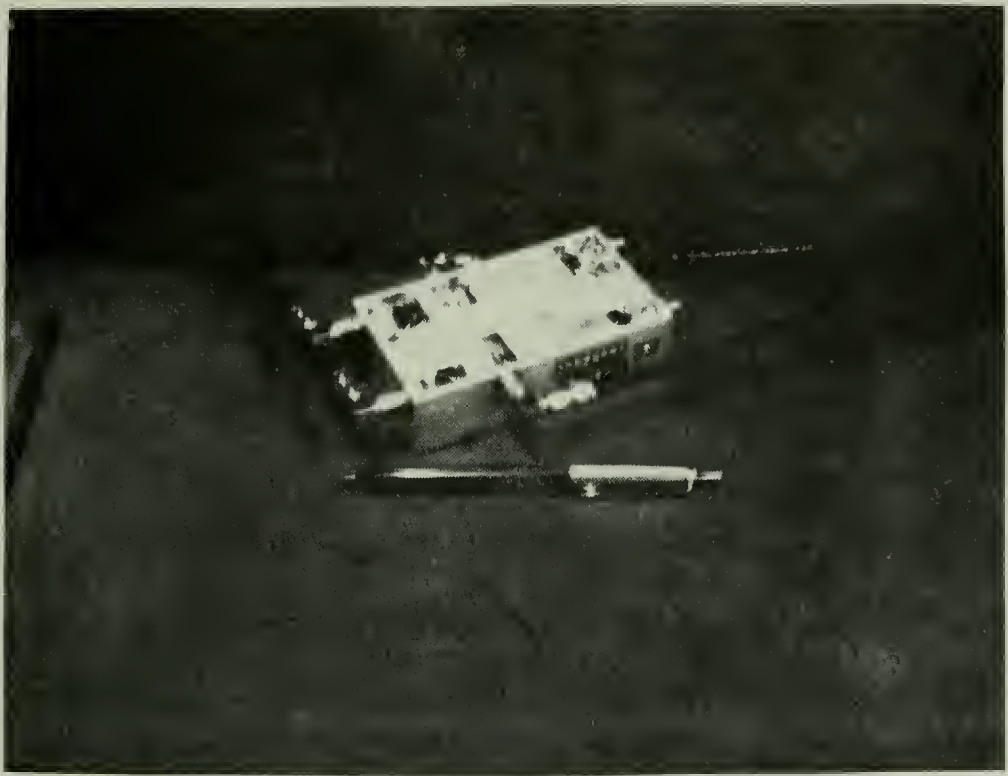

Figure 26. ASWD Bottom View.

contains a transformer to obtain an impedance match between its respective feedthrough and a 50-okm SMA external feed-through connector, which allows coaxial cable connection to the device.

The quartz substrate measures 3.0 inches by 1.5 inches by 0.1 inches and the 2000 Angstrom thick aluminum deposit occupies a total area of 2.6 inches by 1.5 inches. The de-posited devices are composed of a single, ten-finger pair transducer at each end and sixteen phased coded, ten-finger pair transducers in the center. This phase coding is the

feature that actually implements the Golay sequences. The relative phases of the transducers is shown in Figure 27.

ASWD A:	1	1	1	1	0	1	1	0	1	1	1	1	0	0	0	1	0	1
ASWD B:	0	0	1	0	0	0	1	1	1	0	1	0	0	1	0	0	0	0

Figure 27. ASWD Transducer Phase Coding.

The autocorrelation functions of these codes have major peaks of amplitude 16 with associated time sidelobes. The sum of these functions yield a major correlation peak of amplitude 32 with complete sidelobes cancellation.

This phenomenon was produced by utilizing the impulse responses of the devices, which are depicted in Figure 28.

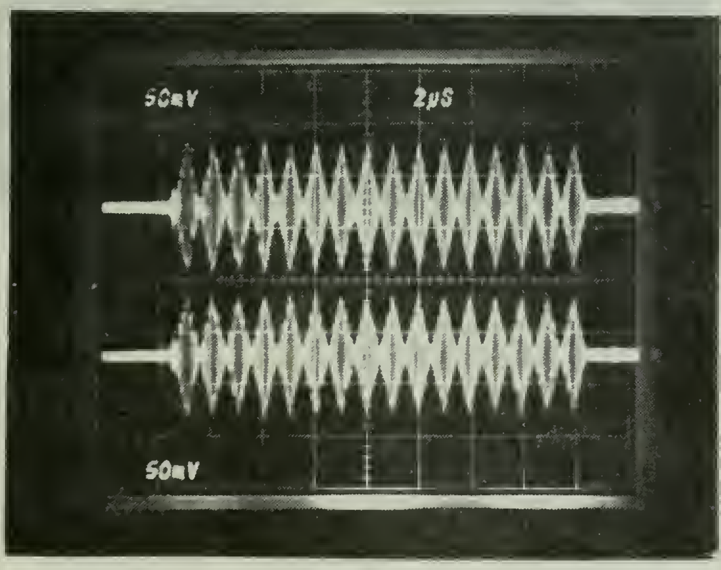

Figure 28. ASWD Impulse Response.

This response was obtained by driving the devices with an electrical pulse of 30 nanosecond width. Each response is comprised of a sequence of 16 pulses of 21.4 megahertz frequency. Each pulse is one microsecond wide, giving the

overall sequence a length of 16 microseconds. The shape of the individual pulses is the result of the convolution process on the substrate of 10-cycle rectangular surface waves with each of the ten-finger pair receiver transducers, which yields individual pulses containing 20 cycles of the carrier frequency with a diamond shaped envelope, as shown in the expanded view of Figure 29. Figure 30 illustrates the phase coding of the impulse responses by summing the waveforms of

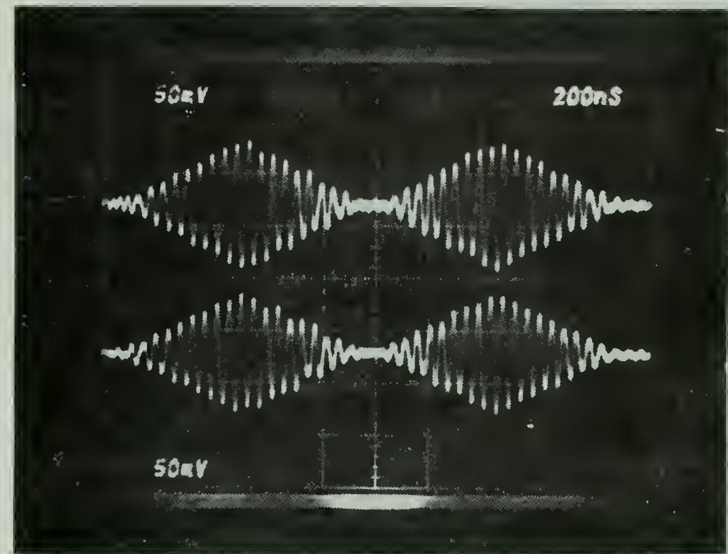

Figure 29. Expanded View of ASWD Impulse Response.

Figure 28. In this case, the first pulses are in phase, and therefore additive, and the second pulses are out of phase and thus tend to cancel each other.

The autocorrelation functions were then obtained by feeding the impulse responses to the appropriate matched devices and the results were as shown in Figure 31. These waveforms clearly depict the dominant center peaks and associated time sidelobes. The sum of these was then taken and the resultant center peak enhancement and sidelobes cancellation that resulted were as illustrated in Figure 32.

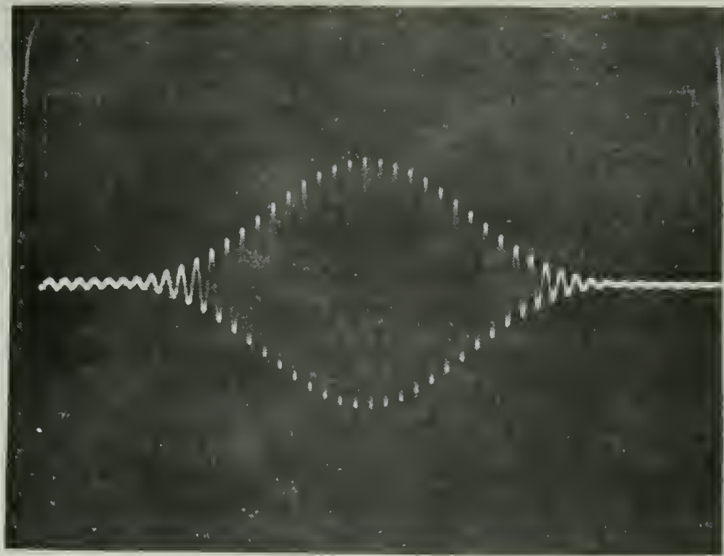

Figure 30. Impulse Response Phase Differences

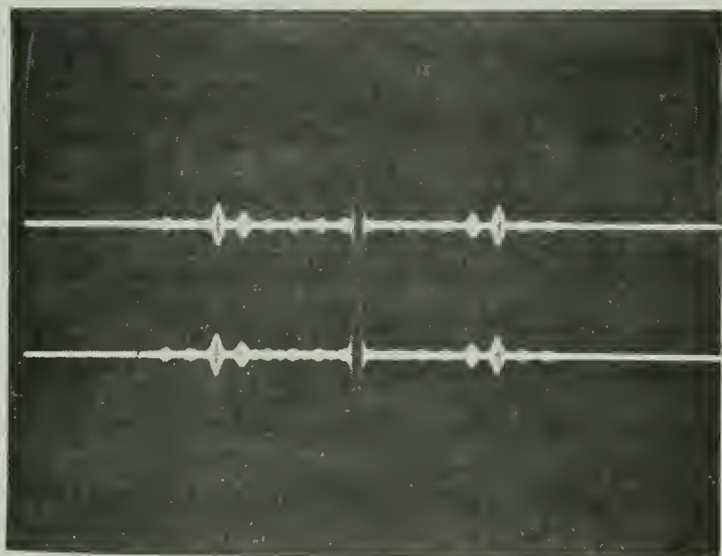

Figure 31. Autocorrelation Functions.

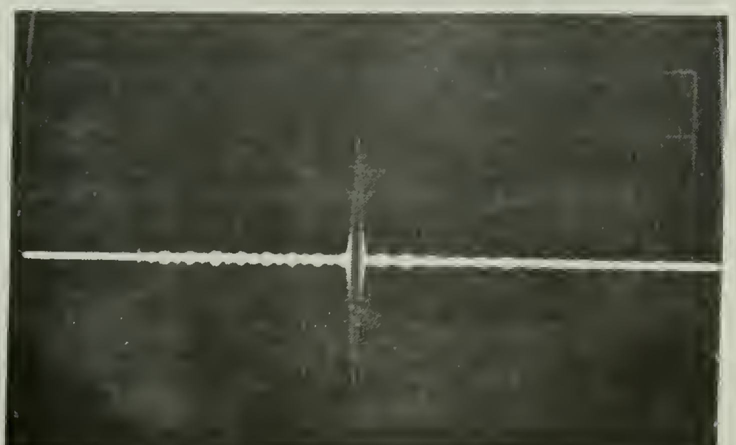

Figure 32. Sum of Autocorrelation Functions.

APPENDIX D: SYNCHRONIZATION CONSIDERATIONS

Frequency Hop and Pseudo Noise code spread spectrum communications systems utilize PN codes in the transmitter to encode the information signal: to spread the bandwidth of the information signal; or more generally, to do a combination of the two. For faithful reproduction of the information signal in the receiver(s) of the system, an identical PN code must be provided to the receiver(s) which is in exact synchronization with the transmitter code. Initial synchronization is accomplished by slewing one of the codes with respect to the other until a usable output indicates synchronization. (Usually the receiver code is slewed through the code pattern; but in some systems the transmitter PN pattern is slewed for searching.)

If the components of the system were stationary with respect to each other, the signal paths were invariant, and the code rates in all of the subsystems of the communications network were precisely identical, initial synchronization would be sufficient to guarantee continuous subsequent synchronization. But since none of these conditions usually prevail, it is necessary that some method of code tracking be provided to compensate for factors which would otherwise tend to break the synchronization of the codes.

Since spread spectrum modulation via ASWD's does not involve the use of a PN code pattern, the sophisticated

circuitry of PN code synchronization is avoided in this latter method. A sacrifice of the PN pattern precludes a system from obtaining some of the advantages usually associated with spread spectrum techniques such as ranging (by measuring code delay time), multiple access (through appropriate code selection), and high degree of communications security (by providing an easy method of code substitutions). In applications where the sacrifice of these features does not effect the requirements of the system, a significant savings in simplicity and space can be gained by employing ASWD's for generation of the spread spectrum waveform.

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